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SAMPLING DIGITAL MULTIPLE CHANNEL SAME FREQUENCY REPEATER STUDY--ETC(U)

JUN 79　W F LAWRENCE, W F KRUSZEWSKI　F33615-78-C-1457

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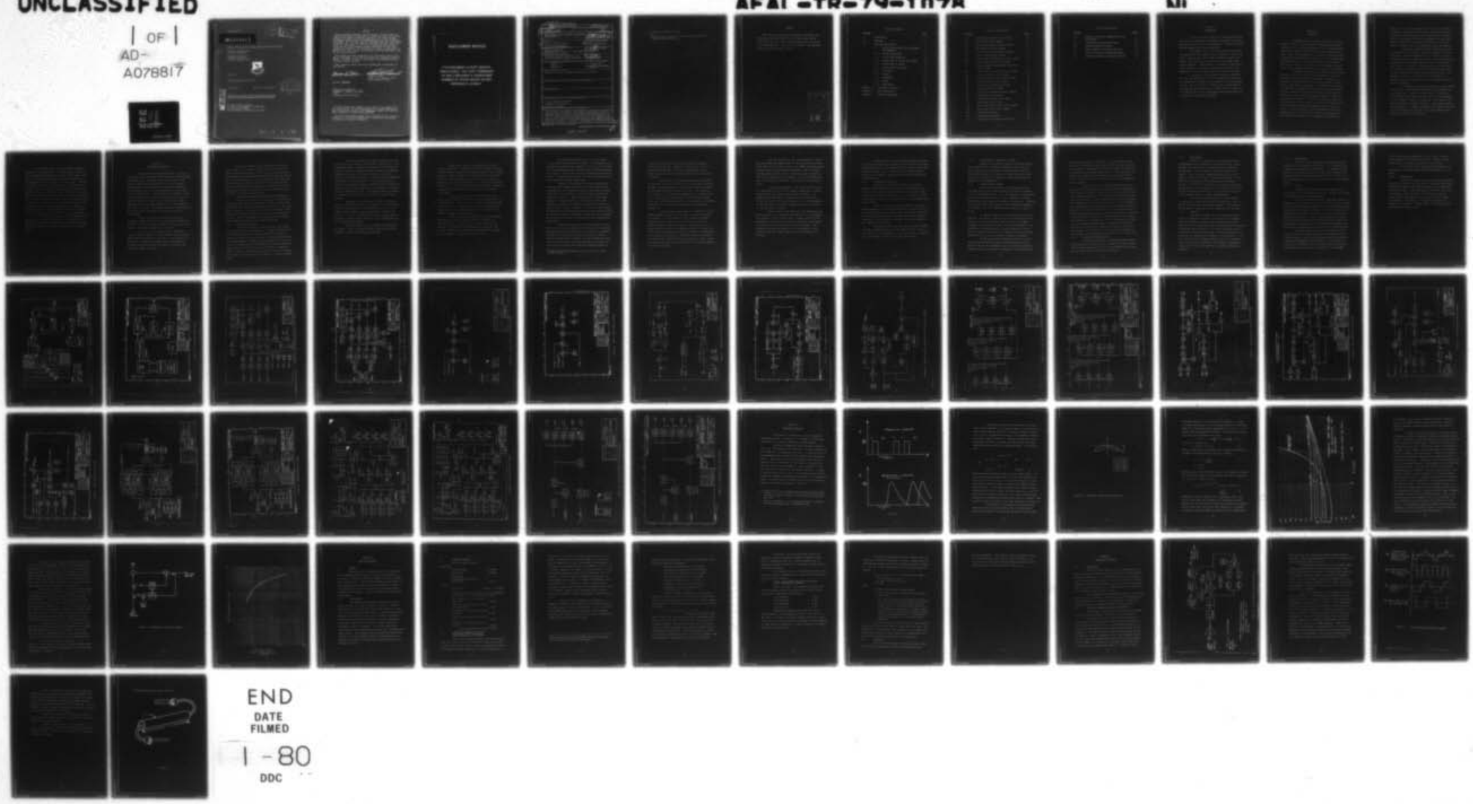

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SAMPLING DIGITAL MULTIPLE CHANNEL SAME FREQUENCY REPEATER STUDY

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JUNE 1979

Technical Report AFAL-TR-79-1078

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The feasibility of modifying an AN/ZRC-1 multiple single frequency repeater was investigated. The present design used an analog sampling scheme to allow the simultaneous repeating of up to four separate analog voice radio links, using only one frequency per link. It was found that this scheme did not lend itself to wide-band digital signals such as those found with the TSEC/KY-58 voice encryption equipment. Instead, a new approach is described which

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would provide a sampling single frequency repeater for wideband digital signals.

PREFACE

This report was prepared by the Bendix Corporation, East Joppa Road, Baltimore, Maryland under USAF Contract Number F33615-78-C-1457, Project No. 7662 and covers work from August 1978 through May 1979. This is a Final Report. The USAF/AFAL project monitor was James J. Nehez.

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SECTION I

INTRODUCTION

This is the final report of the Sampling MSFR Study. As a result of this study, an approach is presented which would provide the Air Force with a new capability; a Secure, Sampling, Multiple Channel Same Frequency Repeater.

The scope of this effort was contractually limited to a sampling single frequency repeater approach. Early in the study, it was apparent that a sampling scheme as used in the AN/ZRC-1 was not directly compatible with the VINSON cryptographic equipment. Presented here is an approach which solves these compatibility problems by modifying the AN/ZRC-1 and developing a small and simple signal conditioning processor.

The Bendix *Communications* Division has developed, over the past several years, a solid foundation of expertise in the marriage of communications equipments to crypto-graphic devices. We are, therefore, confident that our proposed low risk approach would provide the Air Force with the capability of extending UHF line-of-sight <u>secure</u> communications with only a minimum of operational constraints.

SECTION II

OVERVIEW

The MSFR scenario considered here was as follows. Signals from up to four independent communications networks, operating on four separate frequencies, must be passed through an Airborne repeater in order to extend their communications range. Each network operates in a simplex mode using the same frequency for transmitting and receiving. It was assumed that only AN/ARC-164's in combination with TSEC/KY-58's would be used in each network. In addition, it was assumed, for this study, that a sampling technique must be used in the repeater.

The AN/ZRC-1 repeater was developed under an earlier Air Force contract to satisfy the requirements of the above scenario but with one important exception. The ZRC-1 was designed to handle narrow band voice modulated (AM) carriers with no crypto-graphic protection. The addition of the KY-58 for voice encryption makes the original ZRC-1 concept useless. The ZRC-1 sampled all four network channels at once and then sequentially retransmitted each separate network's sample on the appropriate channel. As long as the network's receivers received samples at fast enough rates, the voice signals could be reconstructed by passing the samples through a low pass filter. On the other hand, the KY-58 equipments are digital encryption devices and as such must maintain continuous bit synchronization throughout each transmission and reception.

Therefore, each network receiver must track its own network's transmitter. Sampling all the network's transmitted signals by the ZRC-1's own clock could not be tolerated since it would destroy the individual clock phase information for each transmitter. Instead, the ZRC-1 must be modified to allow each network's transmit clock to be preserved through the repeater.

Some form of transmit/receive isolation must also be maintained at the repeater. The ZRC-1 solved this by only allowing one repeater transmitter on at a time and no transmitters on during the common receive time. This concept can be preserved for each separate channel so that a given channel would either be receiving or transmitting at any given time. But, since each channel must now be asynchronous with respect to each other channel, special attention must be paid to the diplexer and input filter designs of a secure MSFR. In addition, some minor constraints will be necessary on network channel assignments.

The sampling requirement creates yet another problem for the KY-58. The KY-58 was designed to only recognize full width data bits. A sampling repeater would, by definition, only transmit partial data bits which would seriously degrade, if not completely disrupt, the operation of the receive KY-58. We believe that the simplest solution to this problem would be the development of a simple, inexpensive data conditioner which, as will be shown, could be part of the cable assembly which connects the ARC-164 to the KY-58.

As with any single frequency Airborne repeater, backscatter can be significant. The ZRC-1 used a relatively narrow information bandwidth with respect to the channel bandwidth. This allowed the transmit and receive frequencies at the repeater to be slightly off-set from each other. Therefore, the repeater could still be thought of as a single frequency repeater (since only one channel was used) while reducing the backscatter effects. The KY-58, on the other hand, is a wideband system and, when sampled, requires an even wider bandwidth. Thus, one cannot off-set the transmit and receive frequencies in the repeater. We feel that the only practical method to eliminate backscatter in a secure, sampled MSFR is to require the Airborne platform to be flown at sufficiently high altitudes. As shown in Appendix A, these altitudes are very reasonable.

The next section provides a more detailed technical description of modifications necessary for the AN/ZRC-1. In addition, an approach is presented for the ancillary processor that would provide the proper data conditioning for a receive KY-58.

SECTION III

TECHNICAL DESCRIPTION

The proposed Digital MSFR developed as a result of this study effort is similar to the Analog ZRC-1 MSFR developed by Raytheon from a functional block diagram point of view. However, from an operational point of view, the proposed Digital MSFR is very different from its Analog counterpart. The differences are due to the requirements to relay digital 16K Baud VINSON signals. From an analysis of VINSON's digital signal characteristics, along with the particular signal processing constraints imposed by the transceiver and COMSEC equipment used in the communication link, it became apparent that many of the ZRC-1 Analog MSFR operational characteristics and circuits were incompatible with those required for a Digital MSFR.

The most significant differences between the operational characteristics of the proposed Digital MSFR and the ZRC-1 is that each repeater channel is operationally independent, rather than sequenced as in the ZRC-1.

As a result of the requirement for independent repeater channels capable of relaying VINSON 16K Baud signals, modifications are required to the ZRC-1 in the following circuits areas: transmitter/receiver timing, predemodulation filtering, pre/post detection filtering, power supply, and transmitter output filtering.

Using, as a baseline, the existing ZRC-1 block diagrams provided in the Raytheon Company MSFR final report prepared for AFAL under Contract F 33615-70-C-1774, new functional block diagrams have been prepared describing the proposed Digital MSFR. In the subsections to follow, each module level block diagram describing the proposed Digital MSFR is analyzed and the operational characteristics summarized. A rough order of magnitude estimate for the electrical and mechanical modification to the ZRC-1 is included where appropriate.

3.1 SIMPLIFIED FUNCTIONAL BLOCK DIAGRAMS AND OPERATION

Figures 1A through 10A are simplified functional block diagrams for the proposed Digital MSFR. Figures 1B through 10B are Raytheon ZRC-1 block diagrams shown for reference. The legend shown on Figures 1A through 10A indicate those functional blocks that remain the same, are modified, or are new with respect to the ZRC-1. Each of the Module Level functional block diagrams will be briefly described in the following sections.

3.1.1 DIGITAL MSFR SYSTEM

Referring to Figure 1A, the most significant change on the system level is the addition of 3 transmitter modules and a 4 channel multicoupler. The three additional transmitter modules (one dedicated for each channel) as explained previously, are required because of VINSON signal requirements that will not permit one transmitter to be time shared. The added transmitter modules would most likely be similar to the ZRC-1 transmitter module. However, consideration would have to be given to packaging, electrical isolation and heat transfer to insure an optimum design.

Since the proposed Digital MSFR can operate with one or more receivers operating simultaneously, and with one or more transmitters keyed, a 4 channel multicoupler has been added to reduce the effects of channel degradation caused by transmitter broadband noise and/or Nth order intermodulation products. (Appendix "A" presents an analysis of the performance improvement attributable to the multicoupler.) Also, because the multicoupler filter provides relatively narrow receiver RF selectively during receive intervals, attenuation is provided to the other repeater transmitter carriers. This minimizes receiver degradation caused by the effects of one or more off channel strong carriers.

Physically, a 4 channel multicoupler and combiner, having 1) 4 poles per filter, 2) the capability of withstanding a 20 watt 90% AM carrier, 3) low insertion loss and 4) 7000 channel tuning capability, may be expected to require 2 to 3 cubic ft. of volume. Considering the relatively small size of the ZRC-1, the multicoupler would most likely increase the size of the present ZRC-1 package.

The operation of the Digital MSFR can be described as follows. Since all channels operate identically and independently, only one channel will be described.

Assume that a channel is operational but is not relaying information. A received (on channel) VINSON RF signal is down-converted, detected and processed similarly to RF signals applied to a ZRC-1. However, the predetection IF filter bandwidths and detector frequency responses have been widened to accommodate the 16K Baud VINSON data. The operation of the AFC Squelch circuit is also similar to the ZRC-1 with the transmitter inhibited during AFC acquisition and after finite duration signal fades.

Circuitry, unique to the digital MSFR, has been added after the detector to further process the demodulated 16K and VINSON signal. This circuitry, entitled the sampling demodulator, recovers the digital data and clock signal from the VINSON signal. This digital information is then used as the modulation signal for the transmitter.

A unique operational characteristic of the sampling demodulator is that the digital one/zero information is extracted from the 1st half of the received signal bit and applied to the modulator for retransmission during the last 1/2 Bit. Therefore, the receiver and transmitter duty cycles are the same and of a duration equal to 1/2 bit.

The modulation scheme used in the Digital MSFR is also similar to that used in the ZRC-1. The baseband digital information recovered by the repeater's receiver is first shaped by a cosine squared waveform and then used to amplitude modulate an RF carrier. Wider signal processing bandwidths were required, though, to insure that the carrier would be modulated with most of the spectrum of 16 KBS sampled data.

Because of the repeater's 4-channel asynchronous operation, wide transmit spectrum, and 1/2 Bit duration data retransmission, two repeater and system[1] operational characteristics need to be recognized. These are channel spacing and backscatter rejection. Channel spacing in the Digital MSFR repeater is limited primarily by the attention of Nth order transmitter intermodulation product signal levels relative to desired signal strength. Appendix "B" shows that 5 MHz spacing is needed using a multicoupler having relatively good selectivity. Of course, if frequency management is used, repeater frequencies can be selected so as not to coincide with the intermodulation frequencies.

Backscatter rejection of the Digital MSFR is poor due to the fact that the sampled 16 K Baud receive and transmit spectrums overlap and occupy more than 3 channels in bandwidth. Therefore the transmitter/receiver intra-channel frequency diversity scheme used in the ZRC-1 to obtain backscatter rejection can be achieved by flying the repeater at altitudes such that the path loss associated with the backscatter signal is

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[1] Equipment comprising communication link from user to user through repeater.

10 dB or so greater than the LOS path between the repeater antenna and ground user. Appendix A shows that at 5,000 ft. repeater altitude, about a 4 mi. LOS communication link is possible and at 50,000 ft. altitude greater than 100 miles can be achieved.

Three important system characteristics also need to be recognized. The first is that an ancillary signal processing equipment is needed between the transceivers cipher text output and the COMSEC equipments digital data input. The purpose of the processor, is to stretch the demodulated 1/2 bit data pulses to full bit duration compatible with the VINSON COMSEC equipments signal processing requirements. The processor is described electrically and a mechanical configuration is proposed in Appendix "C".

A second system characteristic is whether or not the COMSEC equipment's time delay (TD) mode is required when using the repeater. A finite duration of the transmitting COMSEC equipment's phasing signal must be received for receiving COMSEC equipment to function properly. The duration of the phasing signal inputted to the COMSEC equipment is a function of the following link equipment variables: transmitting COMSEC equipments phasing signal duration, transceiver receive to transmit turn-around time, repeater AFC and sync acquisition time, receive transceivers AGC attack and settling time and ancillary processor sync acquisition time.

When the time taken by the link equipments to change mode and process the VINSON signal is less than the total phasing signal duration by an amount equal to the COMSEC equipment minimum received phasing signal duration requirement, then the normal (minimum phasing signal duration) mode may be used. However, if the latter inequality is not satisfied, then the time delay (TD) mode is required which provides a significantly longer phasing signal.

A third characteristic is the Bit Error Rate degradation associated with the 1/2 Bit sampling scheme used to recover digital data in the repeater and ancillary processor. The repeater's BER degradation is equivalent to that incurred if the RF input signal level were reduced by 3 dB. This degradation is incurred because the sampling demodulator makes use of only 1/2 the available Bit energy.

There is very little additional link degradation due to the presence of the ancillary processor when it is processing the repeater's 1/2 Bit duration data. However, there is a BER degradation incurred when the ancillary processor is processing data from another network user not using the repeater. The degradation is again equivalent to a 3 dB reduction in input signal-to-noise ratio and is due to the 1/2 Bit sampling of the data.

Elimination of the latter degradation may be possible by incorporating circuitry to sense the Bit duration of the input data and bypass the 1/2 Bit sampling circuitry when full Bit data is being received. But, it is felt that the excellent SNR's enjoyed by the repeater's improved LOS should more than off-set this degradation.

3.1.2 DIGITAL MSFR FRONT END

Referring to Figure 2A, the ZRC-1 input hybrid splitters and limiters have been deleted. The antenna is connected to each repeater receiver through a multicoupler filter section and a solid state T/R switch. The T/R switch is cycled from transmit to receive in synchronism with each channels unique timing signal.

Three 1 watt drivers have been added to drive the 3 added transmitters. The drivers would most likely be similar to the existing ZRC-1 driver although some consideration would have to be given to thermal effects, receiver/transmitter isolation, packaging and the mechanical requirement to provide a separate R.F. output for each driver.

The switch driver functional block has been modified to permit independent control of each SPDT RF switch. This is consistent with the requirement to permit any channel or channels to be in transmit or receive at any particular time.

3.1.3 DIGITAL MSFR IF SWITCH XTAL FILTER

Referring to Figure 3A, the most significant modification required was to widen the bandwidth of XTAL filter FL1 from 50 kHz as used on the ZRC-1 to approximately 75-80 kHz. This widened bandwidth is required to insure that most of the energy of the raised cosine sampled 16 KBS received signal is presented to the demodulator.

3.1.4 DIGITAL MSFR IF SWITCH

Referring to Figure 4A, and following the signal flow path from Q1, the first functional block that is modified is the 6 kHz bandpass filter. The 5 kHz bandwidth is appropriate for processing analog signals to 3 kHz. However, the Digital MSFR must process raised cosine sampled 16 K Baud data. Therefore, for the same reason, the crystal filter bandwidth was widened in the IF switch, the 6 kHz bandwidth must be widened to about 75-80 kHz.

It should be noted that the 45 kHz BPF has not been modified. The reason is that the circuitry associated with the filter is only used during AFC acquisition while the COMSEC equipment is outputting 16 K Baud 1-0 phasing information. The spectrum of the phasing signal, including repeater and transceiver frequency, off-sets is still contained within the 45 kHz bandwidth of the filter, therefore, no change is necessary.

Following the signal through Q12 and Q13, the next two functional blocks that need minor modifications are the detector and the 300 Hz bandpass filter. The detector frequency response needs to be widened to accommodate the spectrum of the raised cosine sampled 16 k Baud data. The 300 Hz filter, used as part of the singal-to-noise signal presence circuit, may need to be narrowed. Narrowing the filter would reduce the effects of low frequency 16 K Baud data components (< 300 Hz) modulating the signal to the detector and comparator that follows the 300 Hz filter. This would prevent the generation of false transmitter inhibit commands.

The signal, out of the modified wideband detector, is applied to a new functional block entitled sampling demodulator. Figure 4C shows a detailed circuit level functional Block Diagram of the sampling demodulator. The function of the sampling demodulator is three-fold. First, a clock recovery circuit determines when a VINSON signal is being received with a 16 K Baud preamble. The clock recovery circuit phase locks to the incoming signal and maintains this lock for the duration of the VINSON transmission. Second, once phase lock occurs, a 16 kHz clock signal is outputted to the transmitter, receiver and timing module to insure proper data sampling and retransmission. Third, simultaneous with phase lock, the incoming received signal is sampled during each 1/3 clock cycle and a decision is made as to whether a one or zero was received. The one or zero is then stored and outputted to the modulator for retransmission during the next 1/2 clock cycle.

3.1.5 DIGITAL MSFR AUDIO SWITCH

Referring to Figure 5A, no changes or modifications are proposed for the audio switch. It should be noted, however, that the audio provided to the front panel speaker is 16 K Baud encrypted data and has no value to an operator.

3.1.6 AFC/SQUELCH

Referring to Figure 6A, the only modifications to the AFC/Squelch block diagram are the deletion of two control commands, the first being the frequency off-set command to the modulator. It is not used in the Digital MSFR for two reasons. First, since the transmitted spectrum out of the repeater is about 75 kHz wide, offsetting the carrier 7.5 kHz would not significantly reduce back-scatter interference. Secondly, since ARC-164's receiving the repeaters signal have pre-detection IF filter bandwidths on the order of 70 kHz, offsetting the carrier would result in less energy being applied to the receivers detector, causing BER degradation.

The second command deleted is the error command. It inhibits all transmitters during AFC acquisition. This command is not used since each repeater channel of the proposed Digital MSFR is independent and not sequenced as in the ZRC-1.

3.1.7 EXCITER

Referring to Figure 7A, three functional blocks need to be modified. The first is FL1, whose 6 kHz bandpass needs to be widened to about 50 kHz to adequately pass the spectrum of the 455 kHz signal amplitude modulated by the digital data pulses from the receiver. The second functional block to be modified is the HI-LO oscillator. The change proposed is to delete on L.O. and shift the frequency of the remaining L.O. 7.5 kHz so that the final transmit frequency off-set is deleted to insure that as much as possible of the repeaters transmitted signal will fall within the ARC-164's receiver pre-detection bandwidth.

3.1.8 TRANSMITTERS

Figure 8A is the functional block diagram for each transmitter of the proposed digital MSFR. It is identical with Figure 8B, the original ZRC-1 transmitter. No changes are anticipated for this module, although as mentioned previously, four transmitters will be required and consideration will have to be given to packaging, RF isolation, heat transfer, and power supply loading.

3.1.9 TIMING

Figure 9A is the functional block diagram of the timing module for the proposed Digital MSFR. Changes are proposed for the Analog and Digital subassemblies.

The major change to the Analog subassembly would be to change the clock frequency from 60 kHz, as used in the ZRC-1, to 16 kHz. The 16 kHz clock would be synchronized with the recovered frequency of the channel's sampling demodulator. This circuitry would be repeated for each channel, in lieu of one circuit common to all channels as in the ZRC-1. Similar cosine-squared wave shaping circuitry would be used, with the exception that the transmit and receive duty cycles would be 0.5 rather than 0.25 as used for the ZRC-1 during recieve and 0.32 or 0.28 for the ZRC-1 during transmit.

The major change to the digital assembly would be to delete the circuitry associated with the sequencing of the repeater channels. New circuitry would be added to provide for independent asynchronous channel operation. The circuitry to inhibit all transmitters during AFC signal acquisition on any channel has been deleted since each channel of the proposed

Digital MSFR operates independently of all others. However, a circuit would be incorporated to inhibit a channel's transmitter until clock recovery and AFC lock has occurred.

The transmitter ALC and sample/hold assembly would be used as is with power output control accomplished as in the ZRC-1.

3.1.10 POWER SUPPLY

Referring to Figure 10A, the only change anticipated would be to modify the 28V high current transmitter power supply to provide the additional current required by the 3 additional transmitter modules. Some consideration would have to be given to packaging, electrical design, and cooling to insure an optimum design compatible with the size and voltage isolation requirements of the repeater. The remaining supplies are assumed to be adequate to supply power to the balance of the modified Digital MSFR circuits.

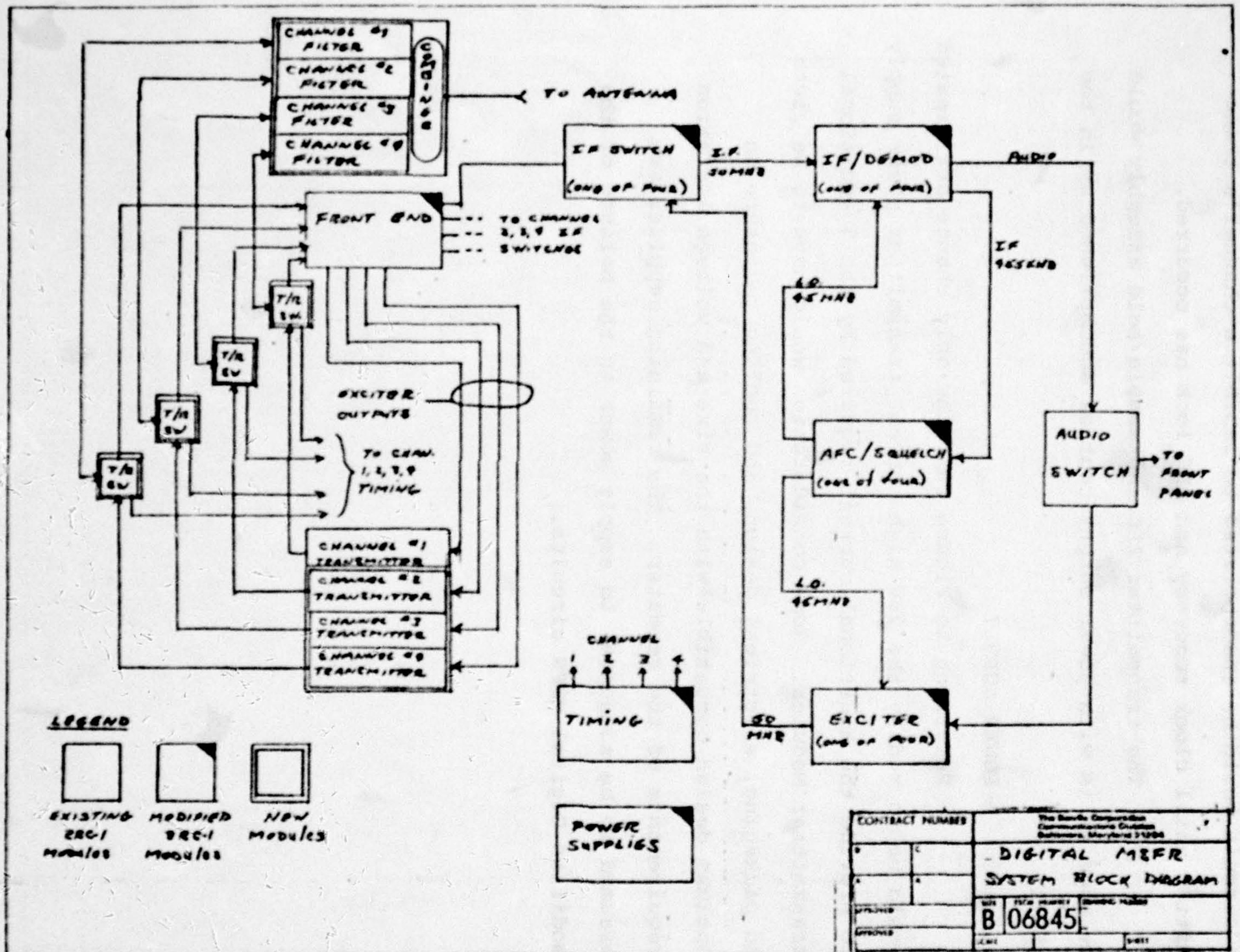

FIGURE 1A. DIGITAL MSFR SYSTEM BLOCK DIAGRAM

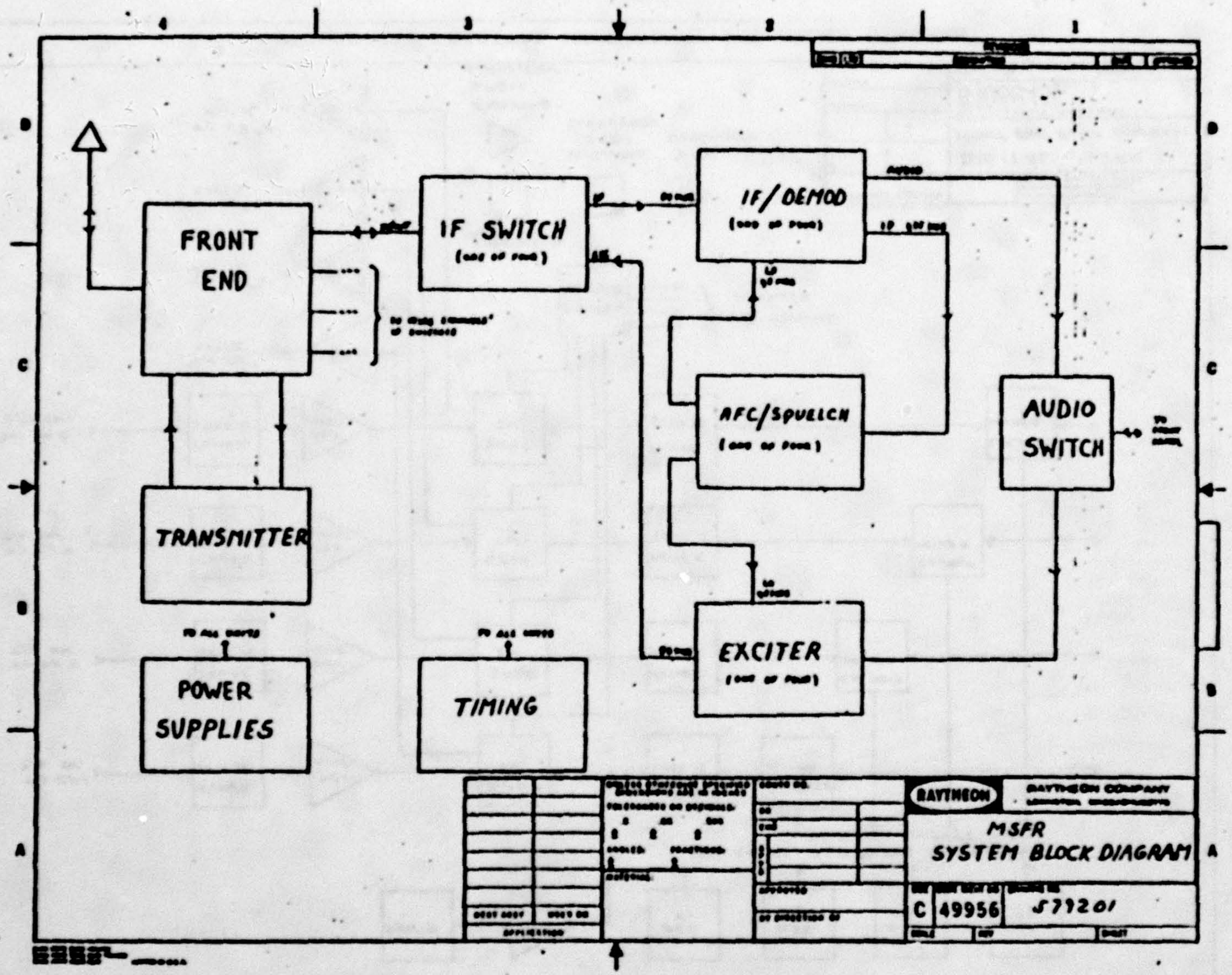

FIGURE 1B. MSFR SYSTEM BLOCK DIAGRAM

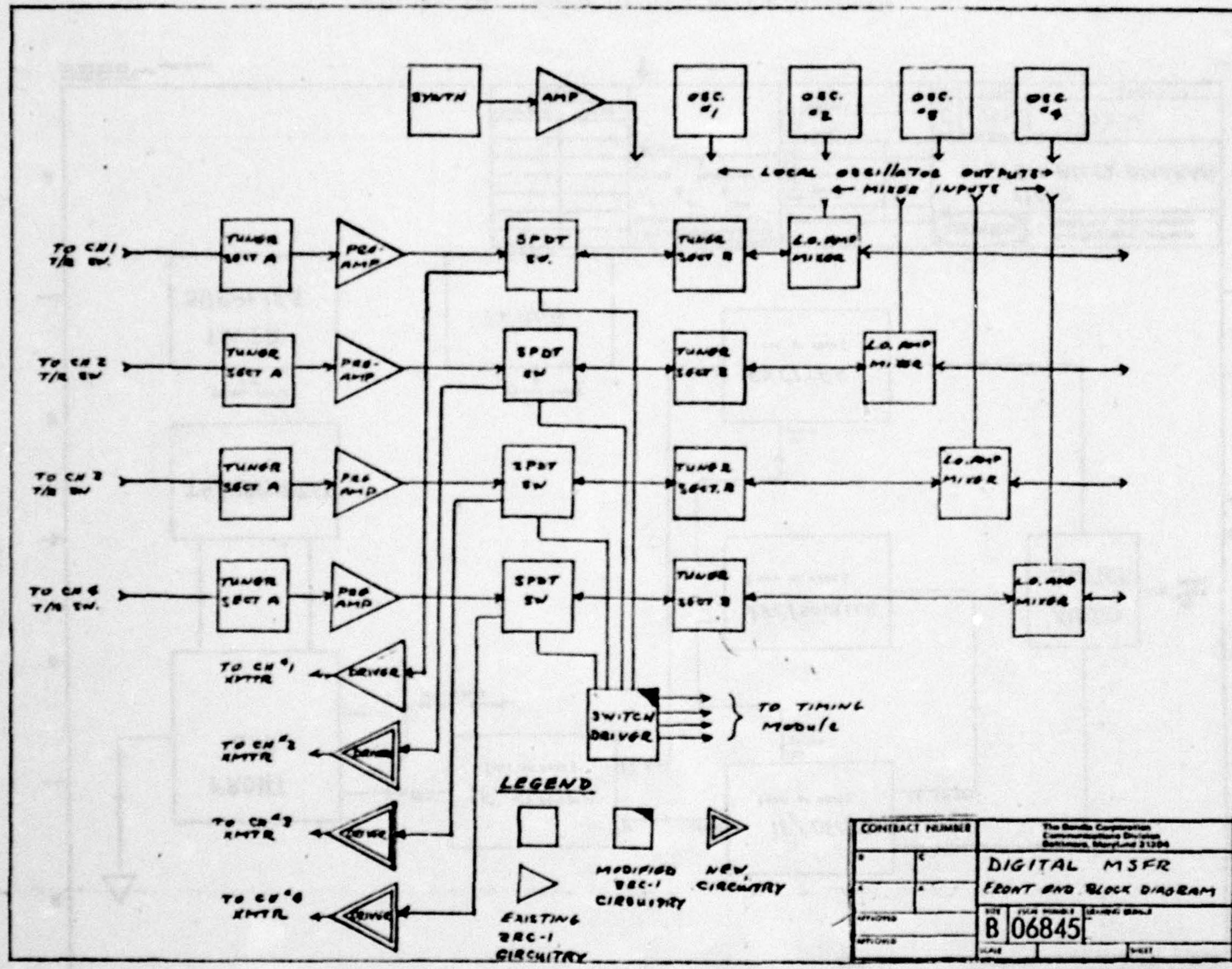

FIGURE 2A. DIGITAL MSFR FRONT END BLOCK DIAGRAM

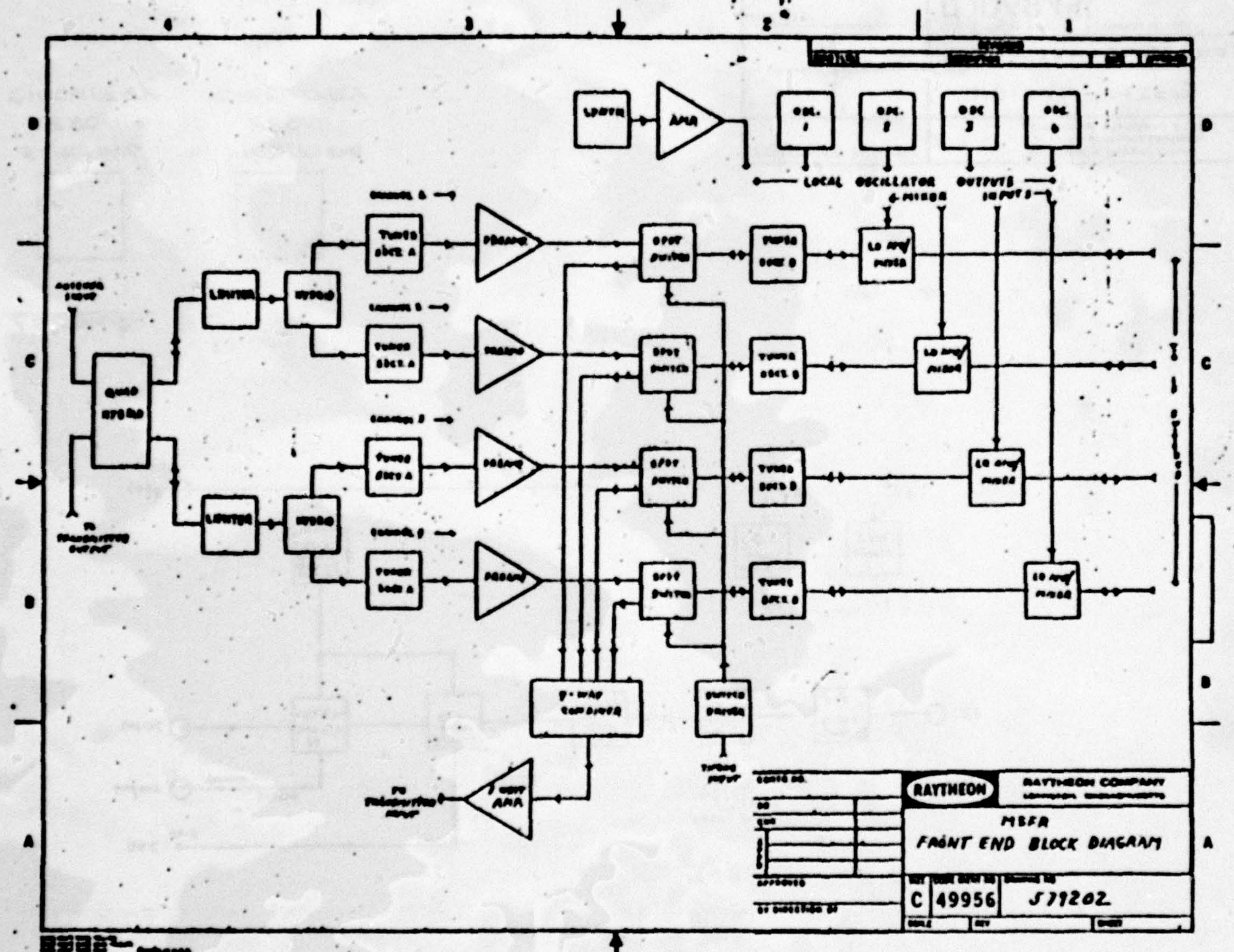

FIGURE 2B. MSFR FRONT END BLOCK DIAGRAM

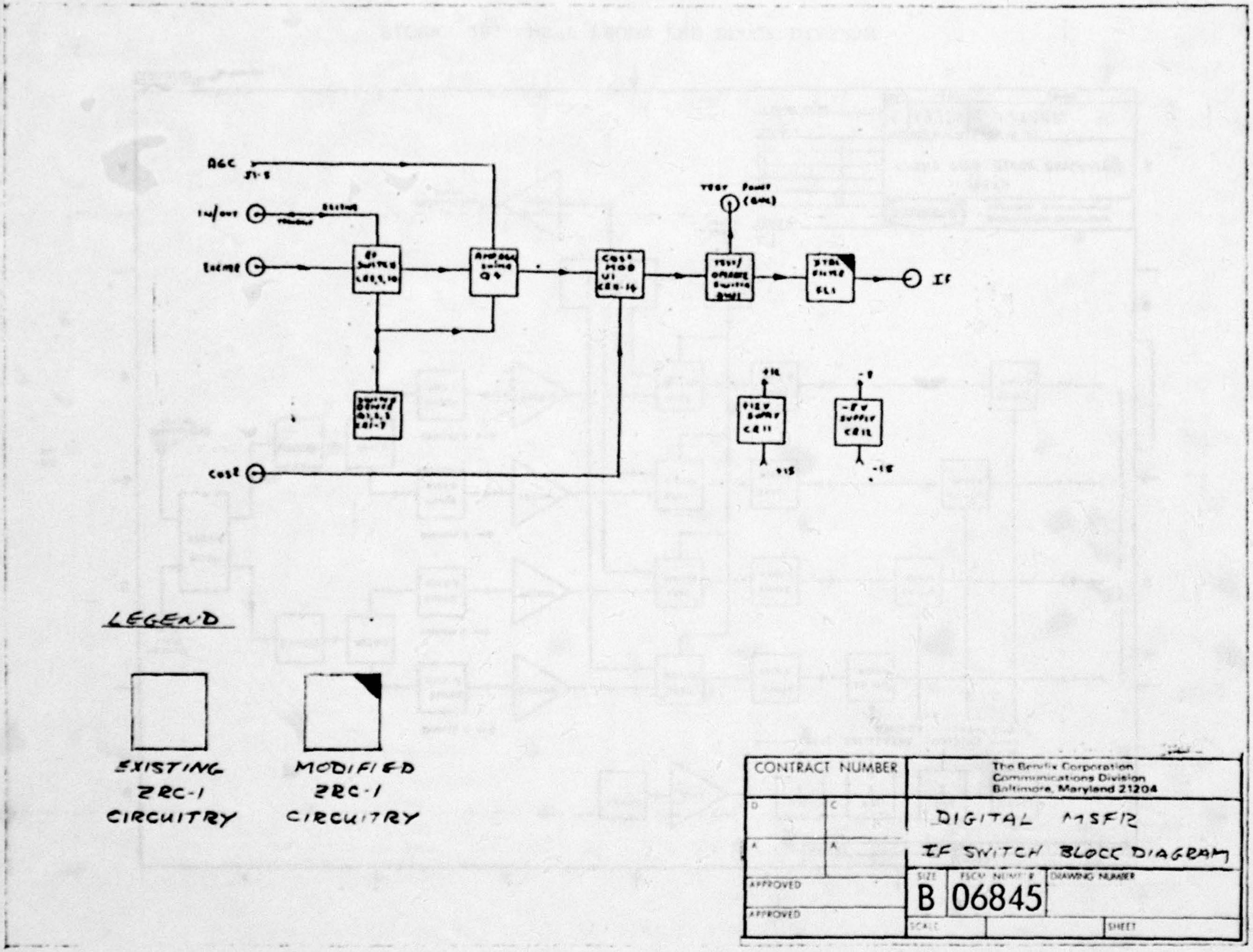

FIGURE 3A. DIGITAL MSFR IF SWITCH BLOCK DIAGRAM

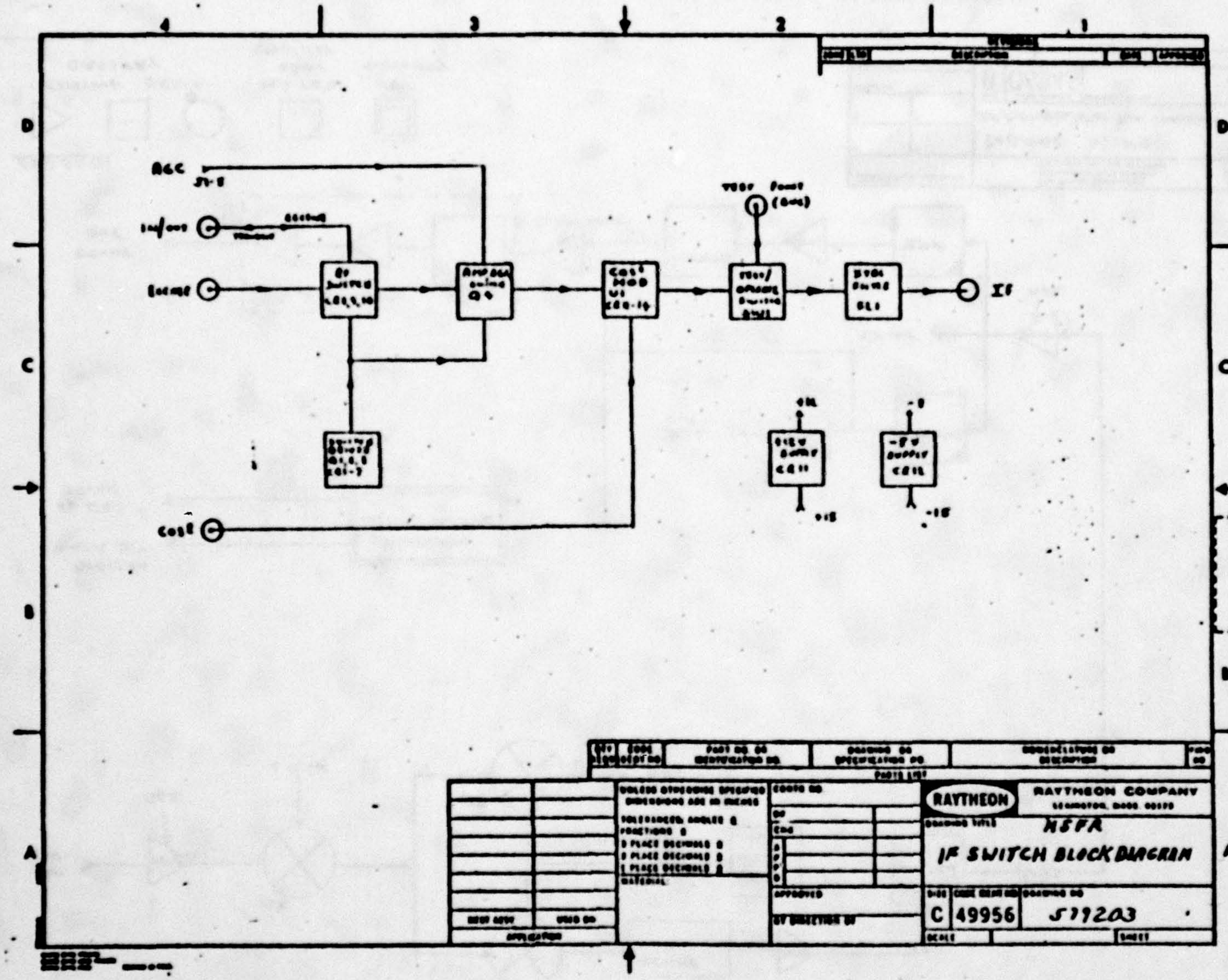

FIGURE 3B. MSFR IF SWITCH BLOCK DIAGRAM

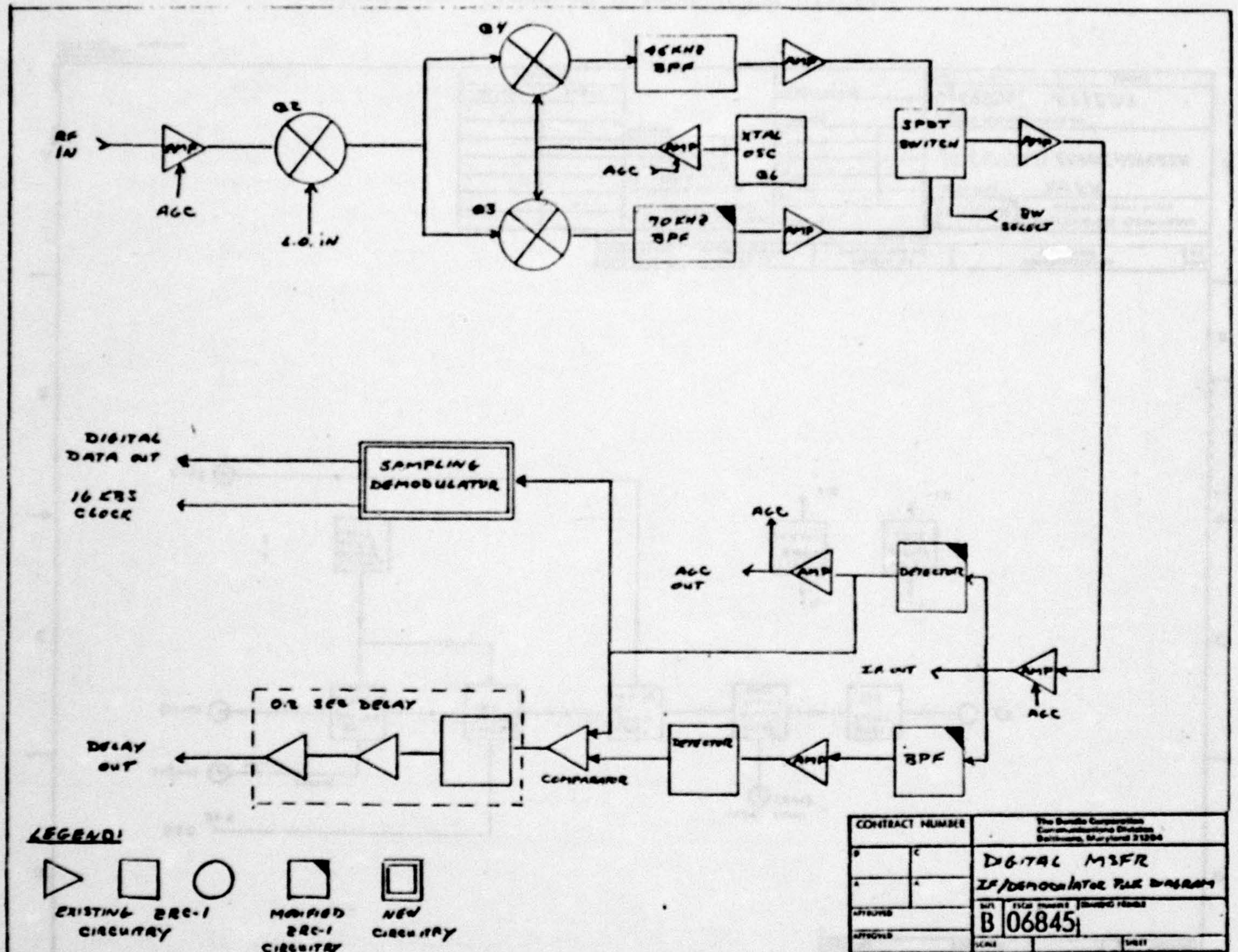

FIGURE 4A. DIGITAL MSFR IF/DEMODULATOR BLOCK DIAGRAM

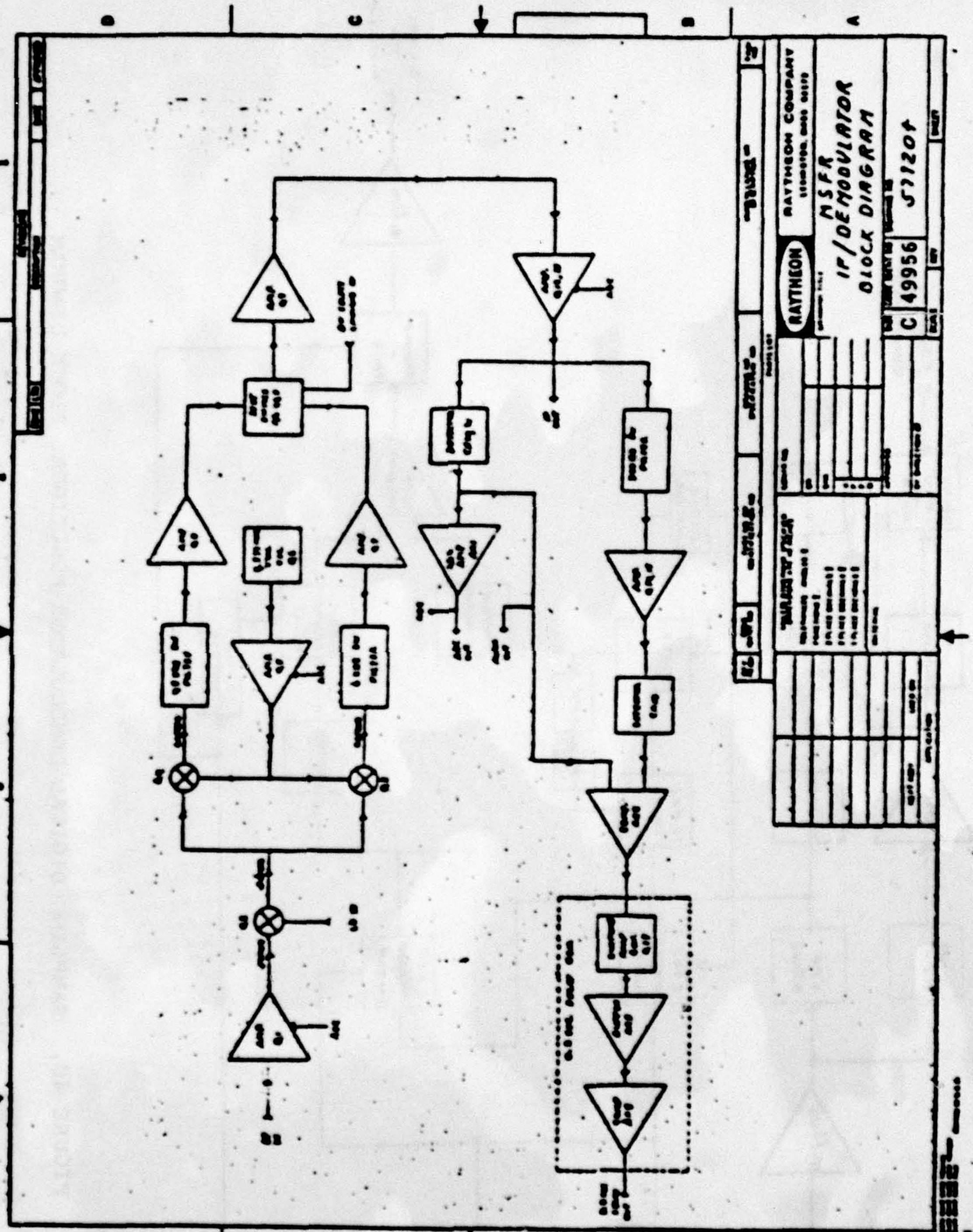

FIGURE 4B. MSFR IF/DEMODULATOR BLOCK DIAGRAM

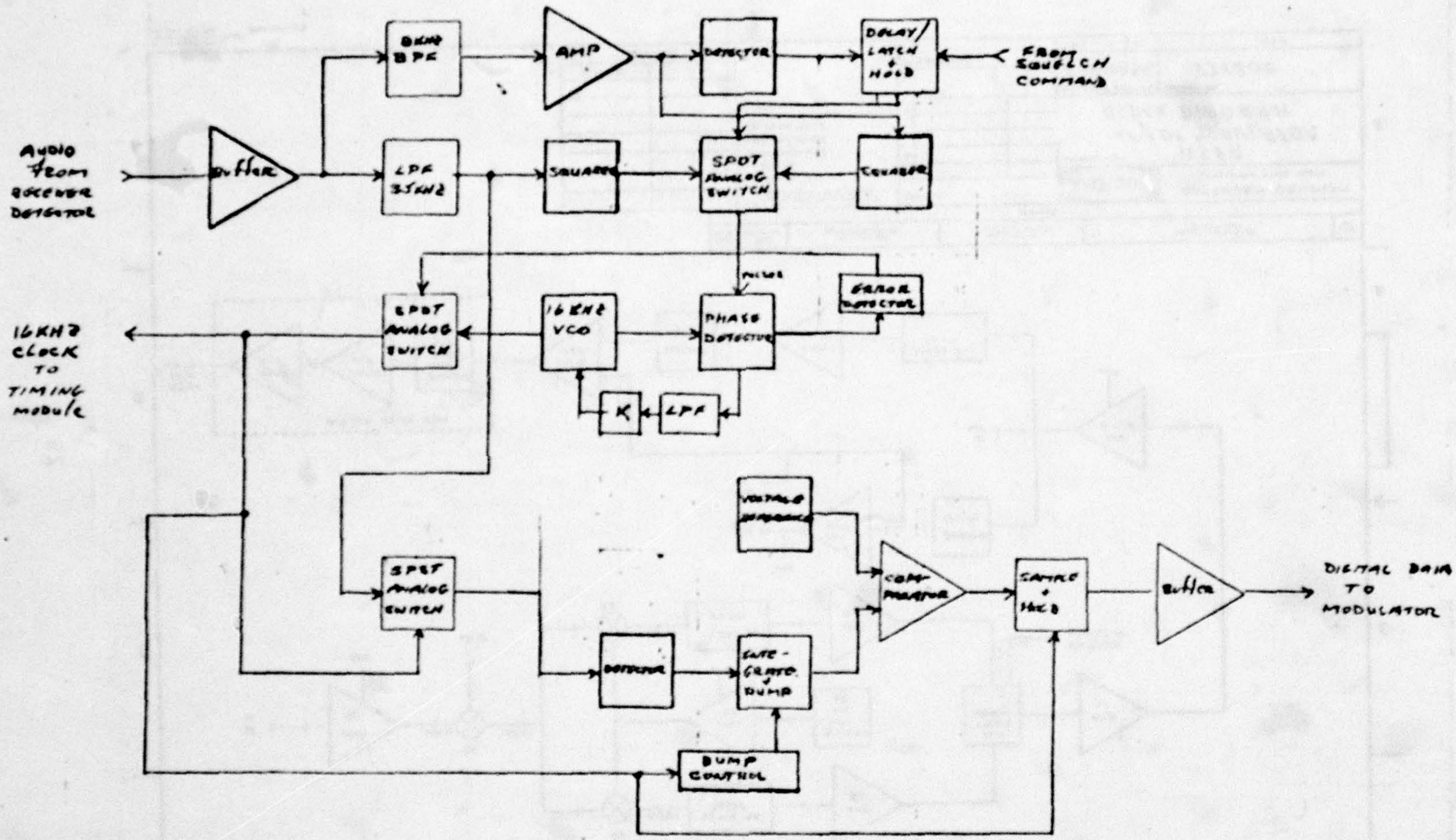

FIGURE 4C. SAMPLING DIGITAL DEMODULATOR FUNCTIONAL BLOCK DIAGRAM

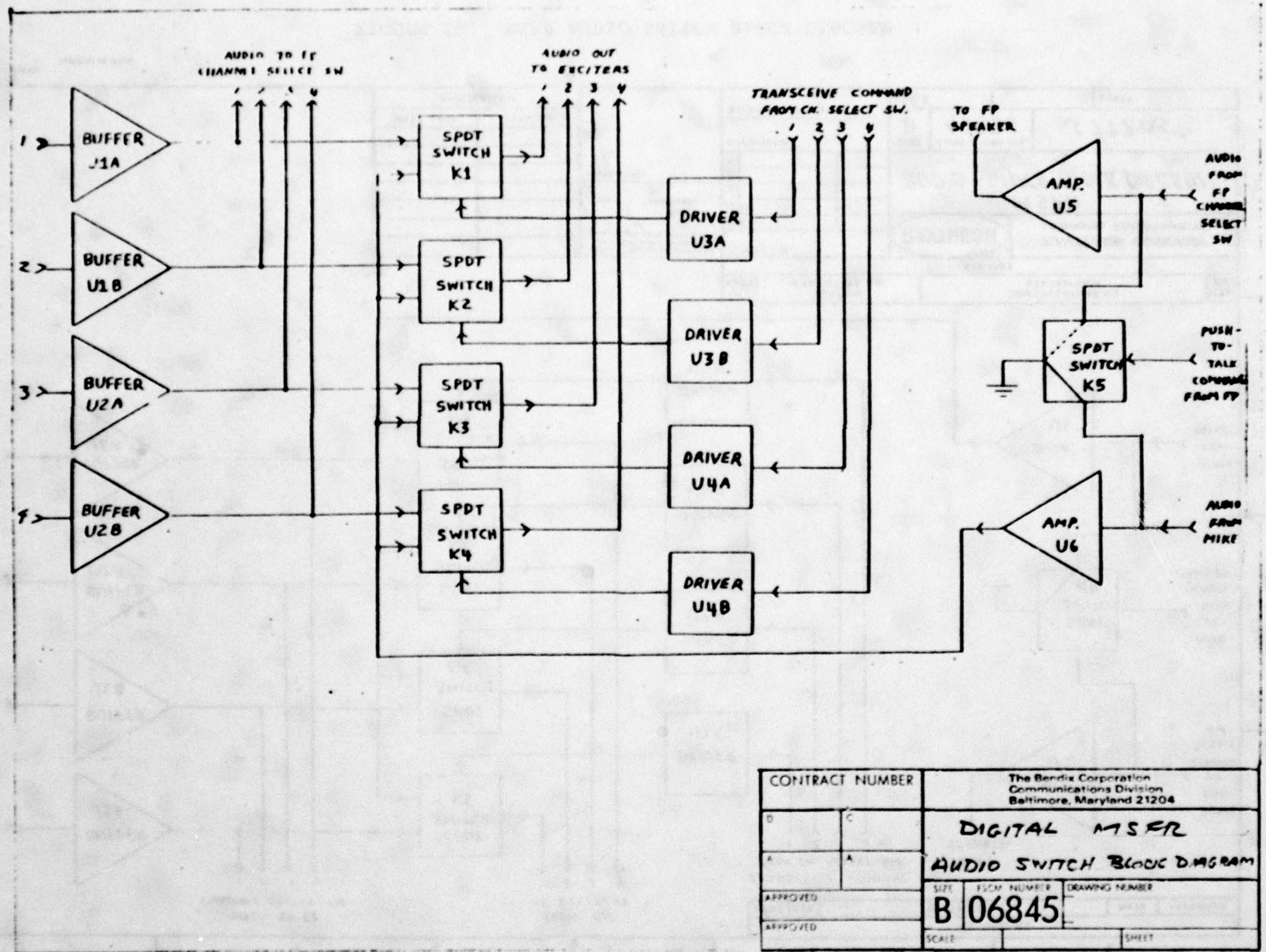

FIGURE 5A. DIGITAL MSFR AUDIO SWITCH BLOCK DIAGRAM

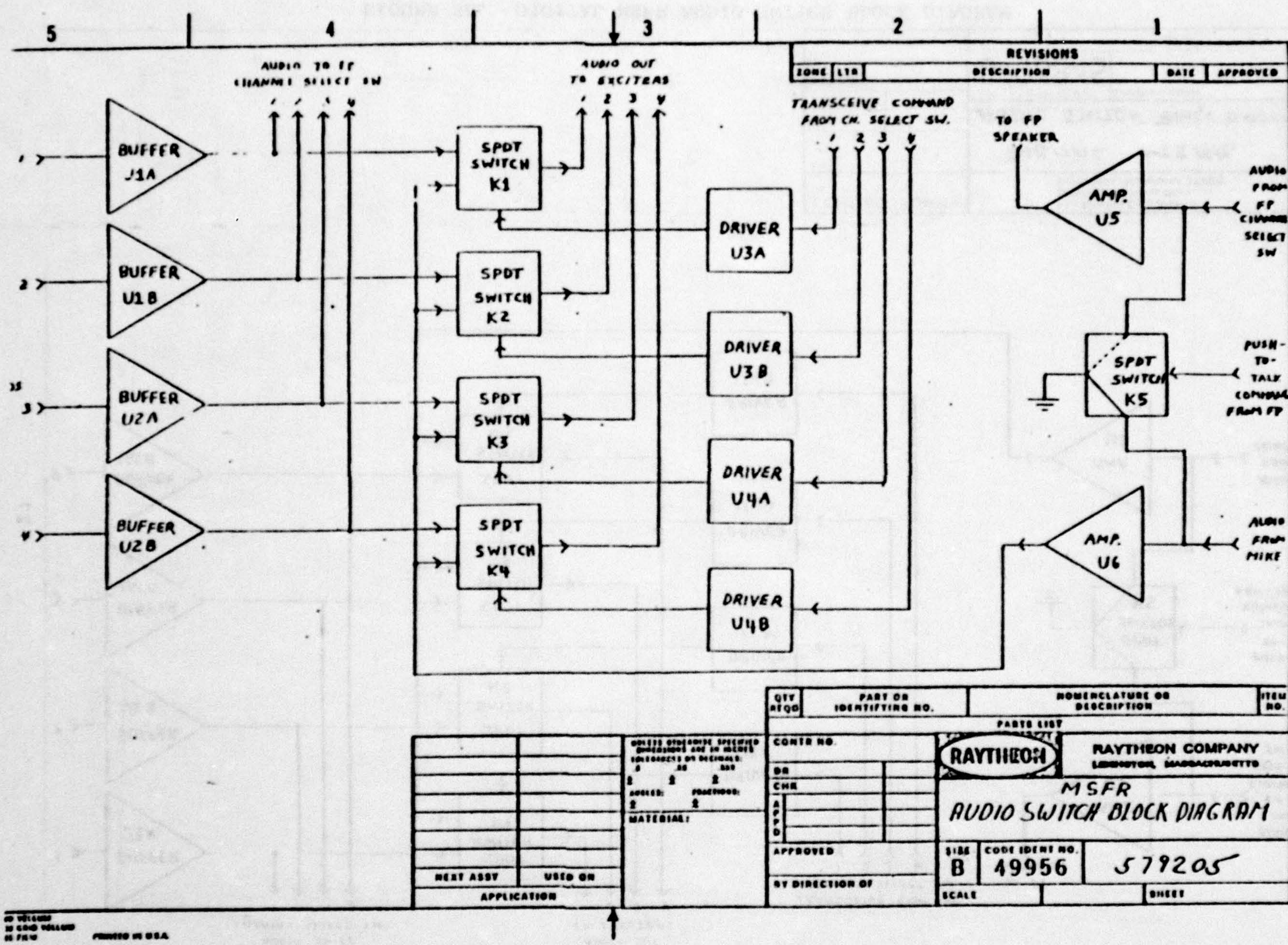

FIGURE 5B. MSFR AUDIO SWITCH BLOCK DIAGRAM

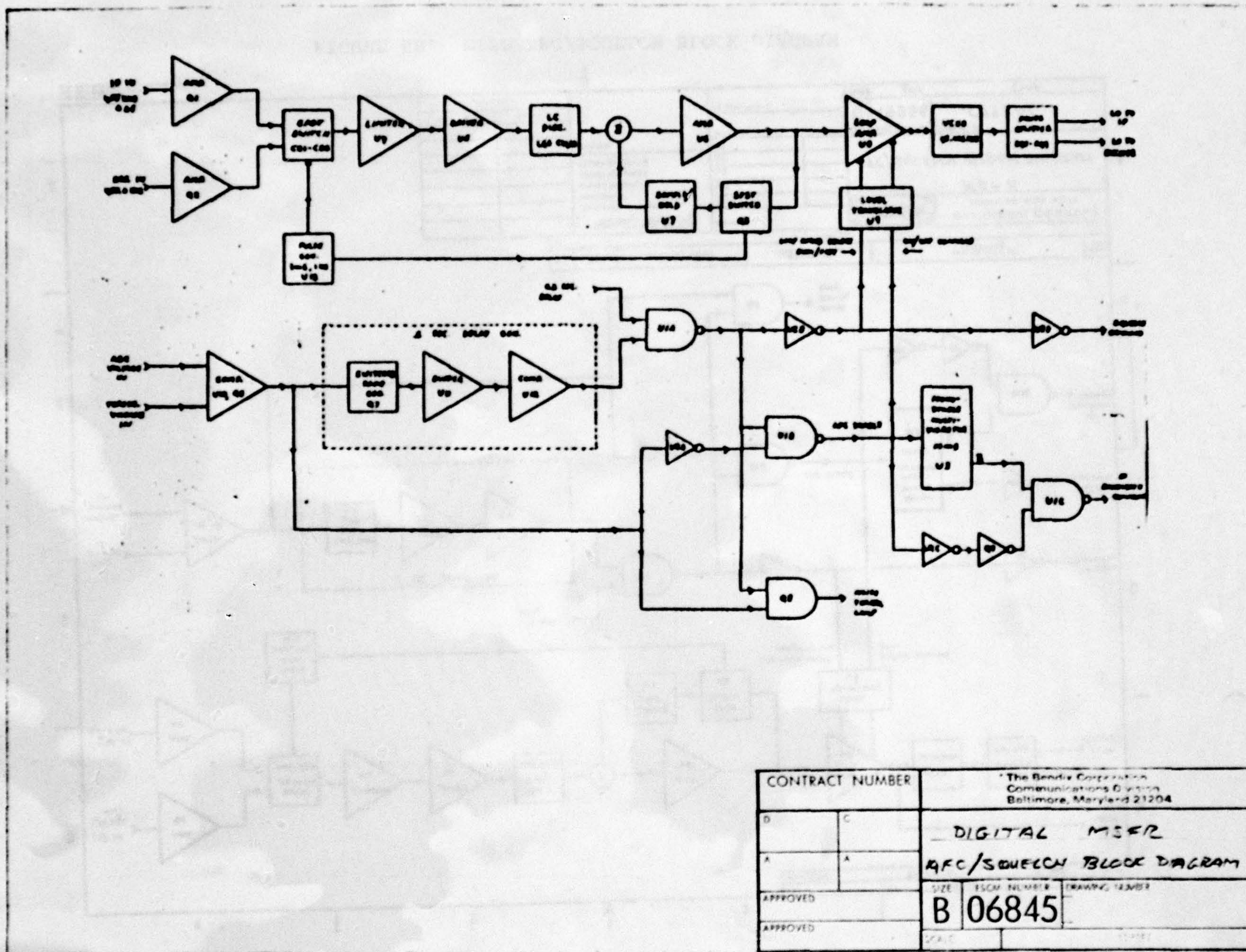

FIGURE 6A. DIGITAL MSFR AFC/SQUELCH BLOCK DIAGRAM

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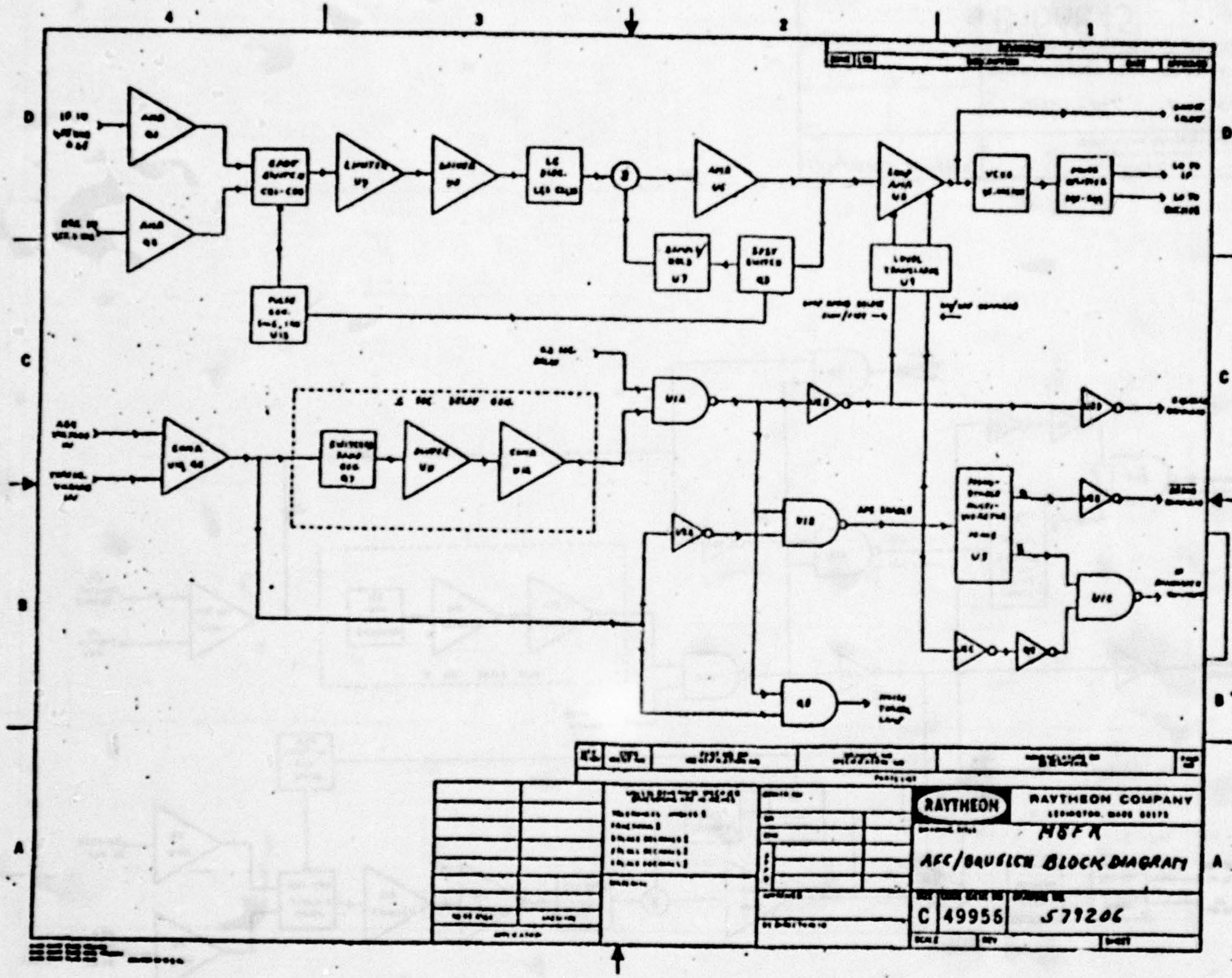


FIGURE 6B. MSFR AFC/SQUELCH BLOCK DIAGRAM

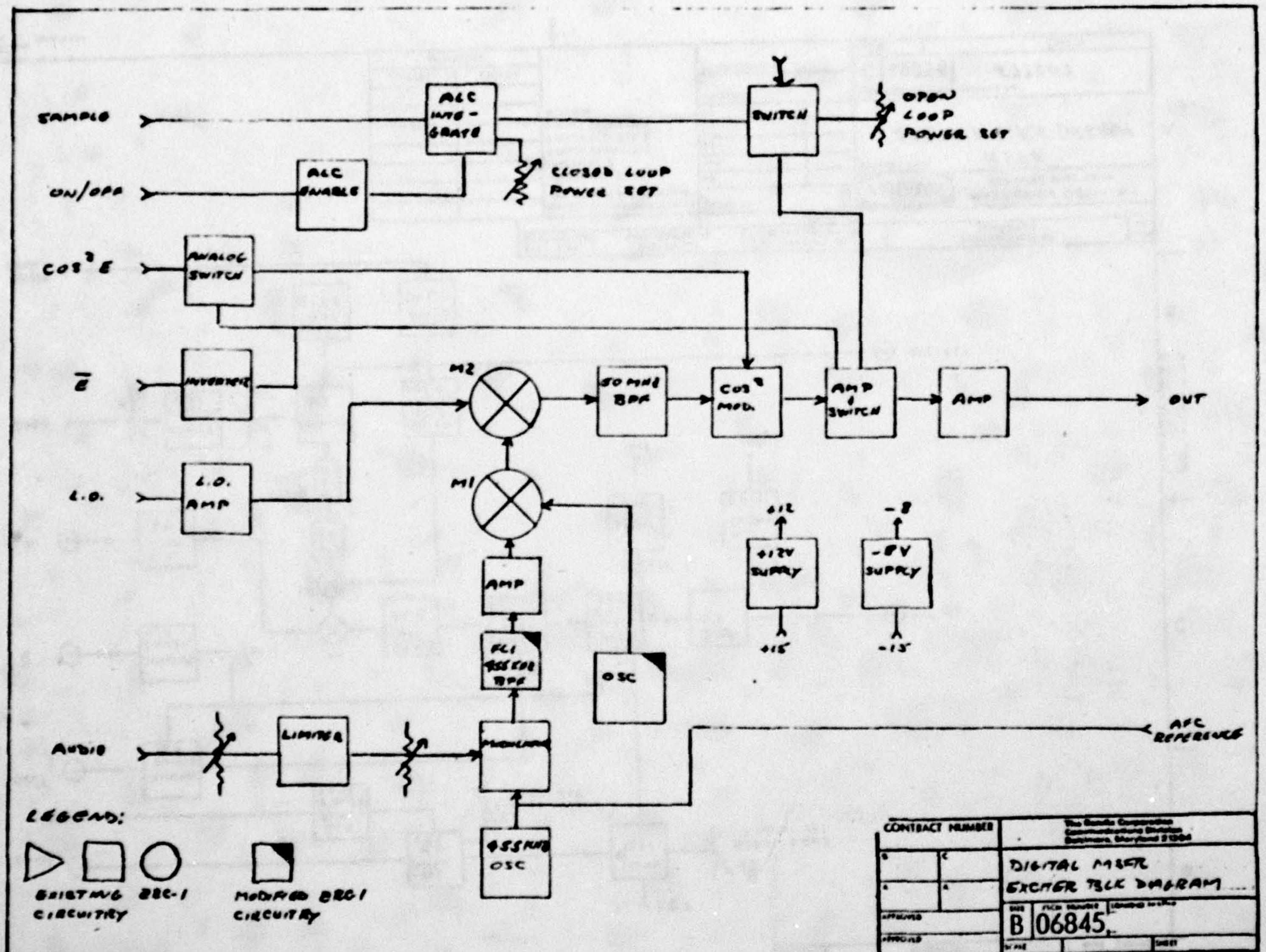


FIGURE 7A. DIGITAL MSFR EXCITER BLK DIAGRAM

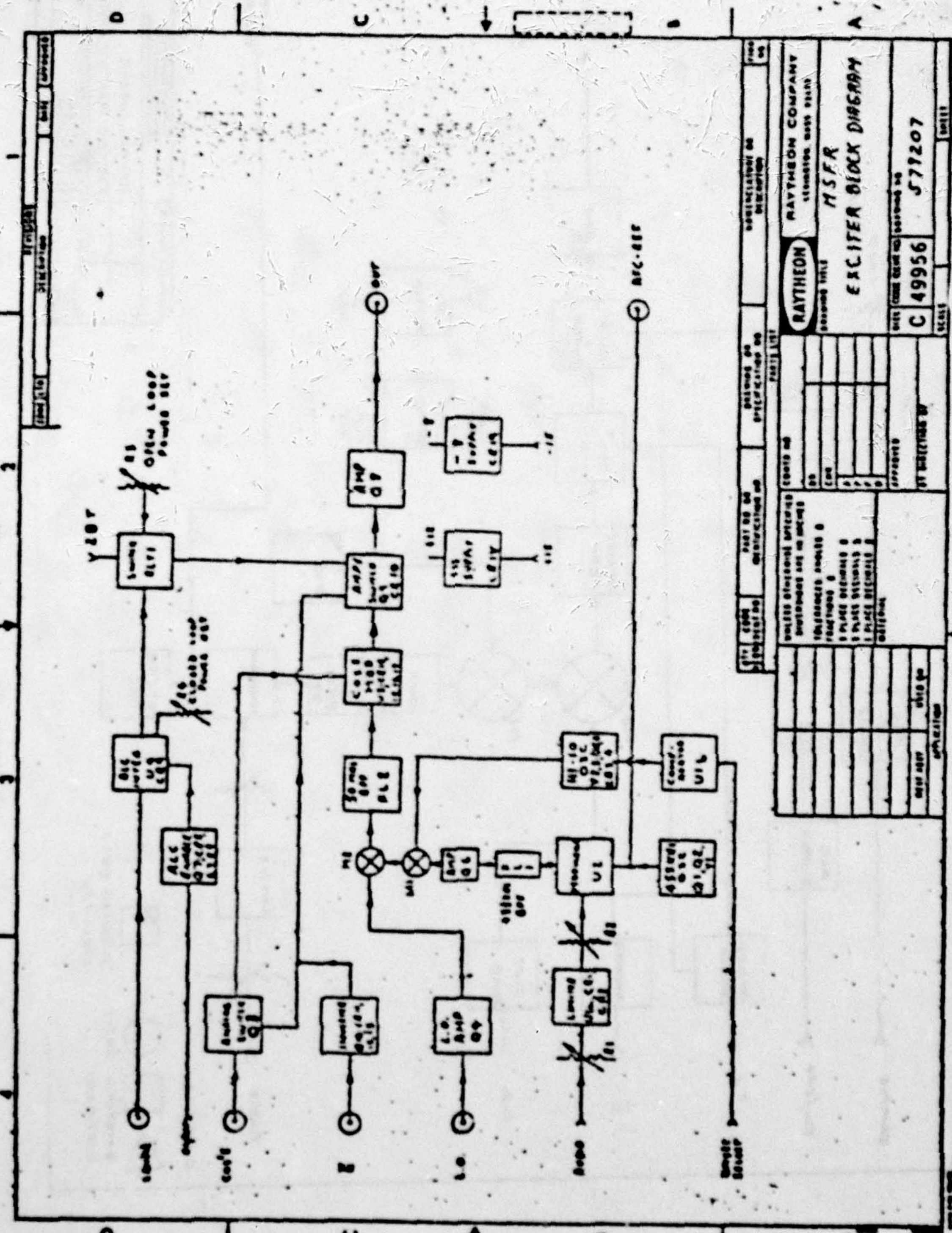


FIGURE 7B. MSFR EXCITER BLOCK DIAGRAM

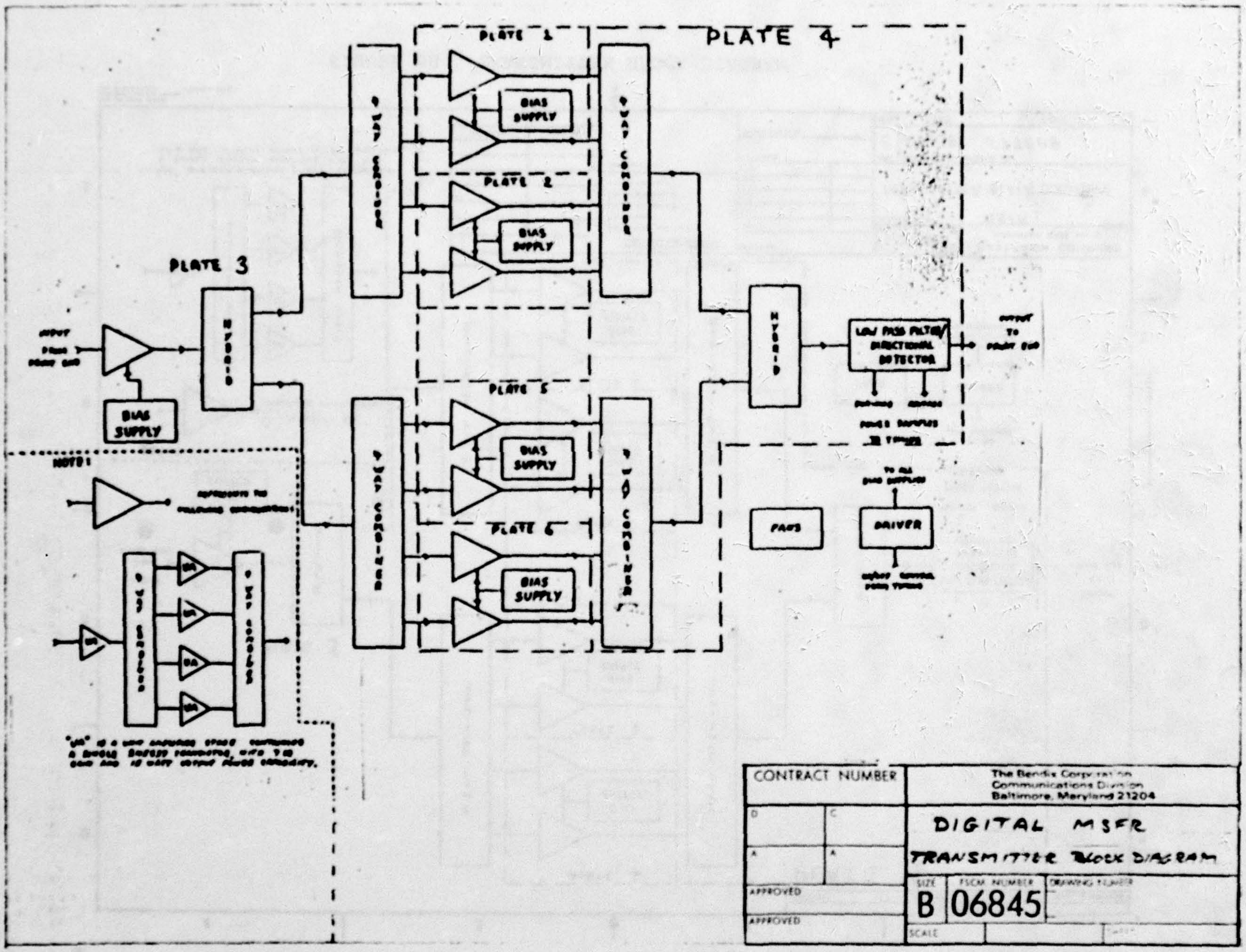


FIGURE 8A. DIGITAL MSFR TRANSMITTER BLOCK DIAGRAM

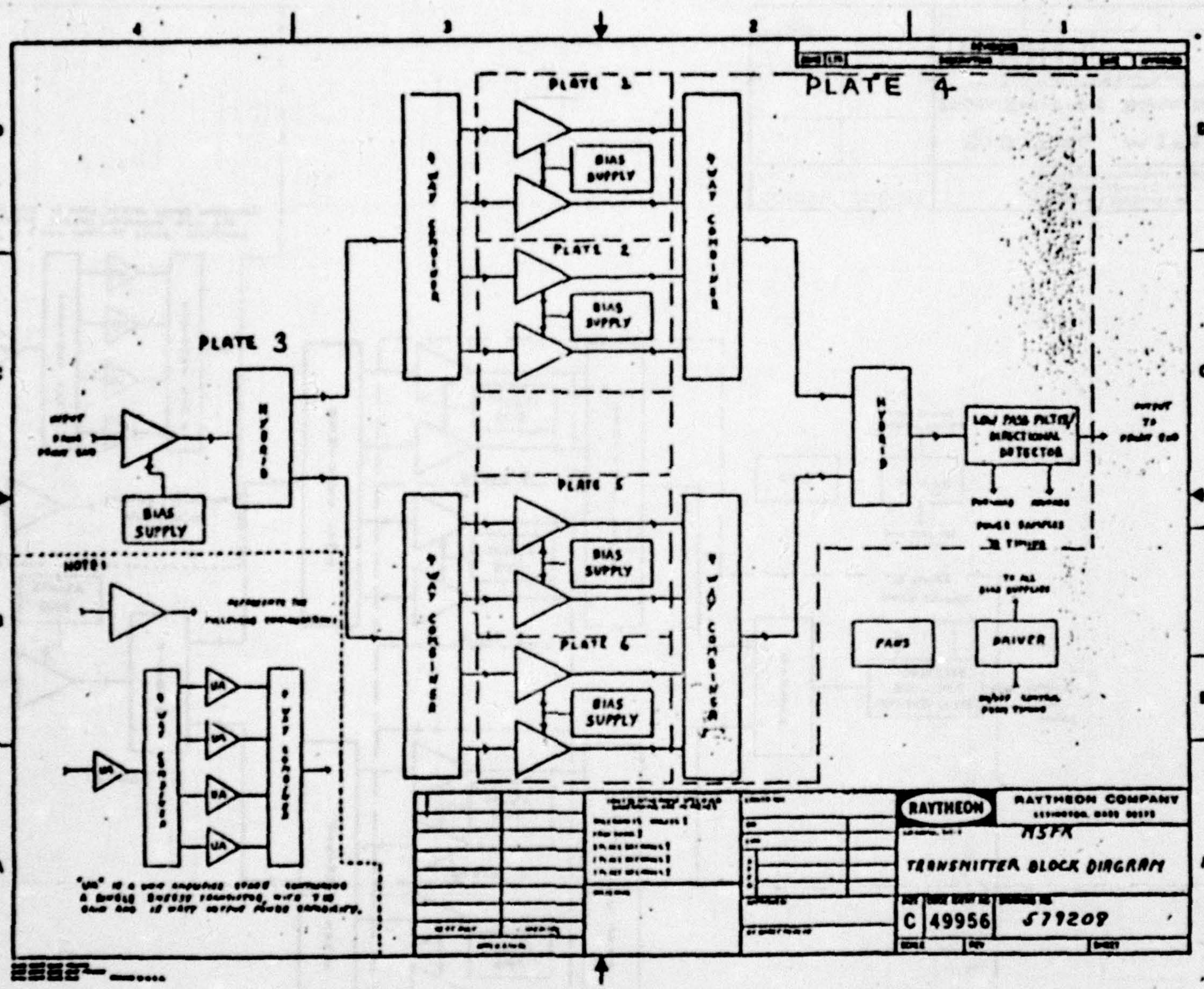


FIGURE 8B. TRANSMITTER BLOCK DIAGRAM

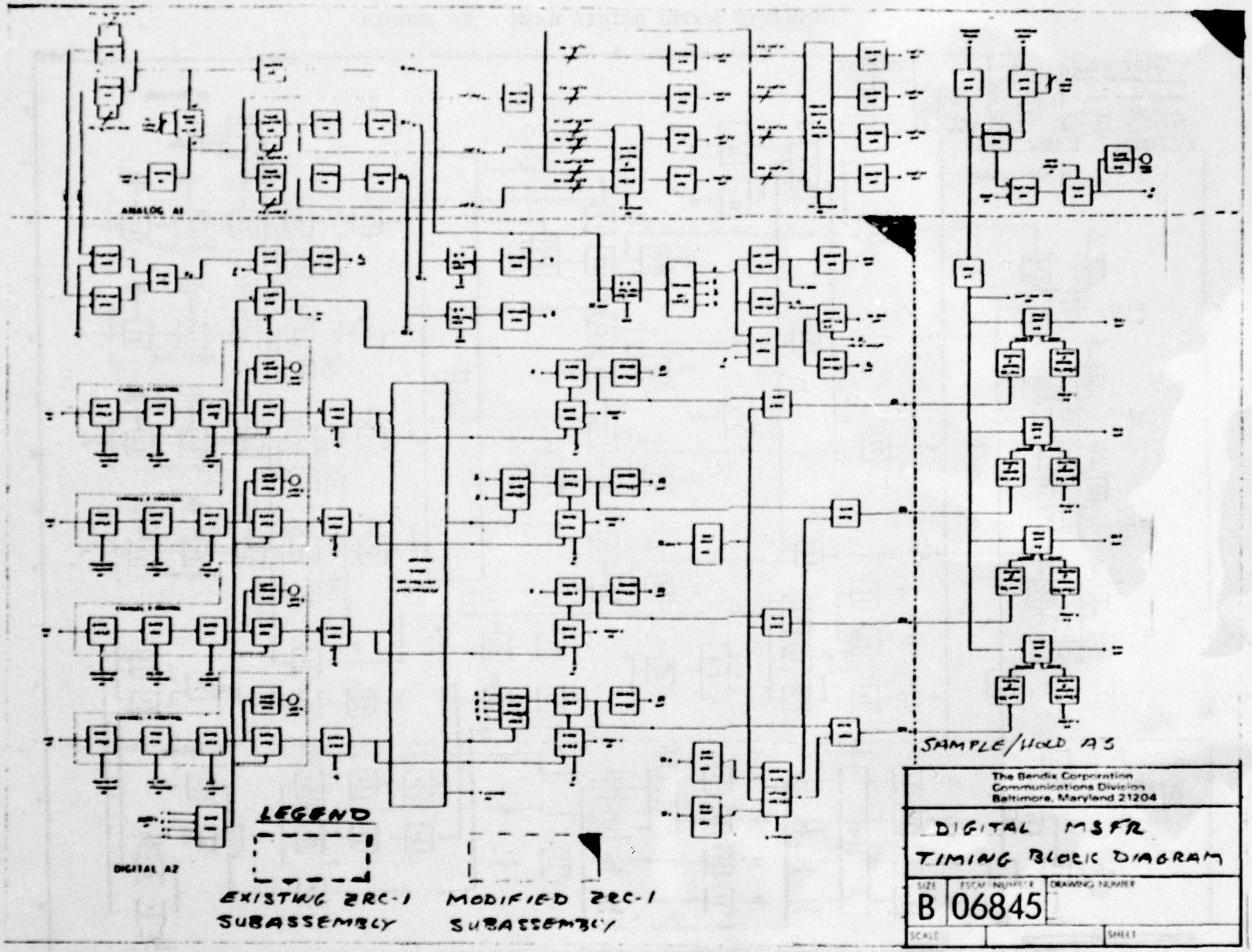


FIGURE 9A. DIGITAL MSFR TIMING BLOCK DIAGRAM

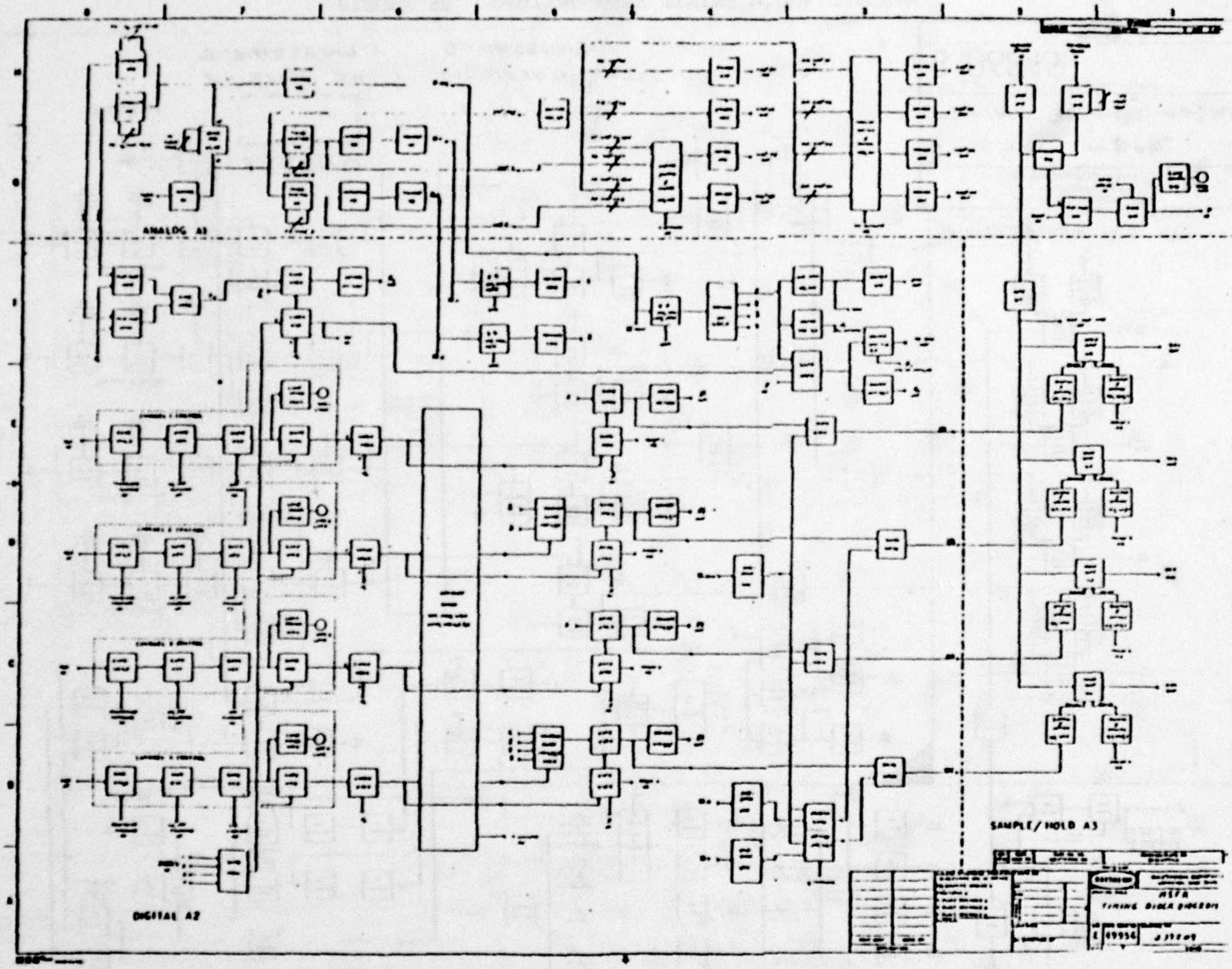


FIGURE 9B. MSFR TIMING BLOCK DIAGRAM

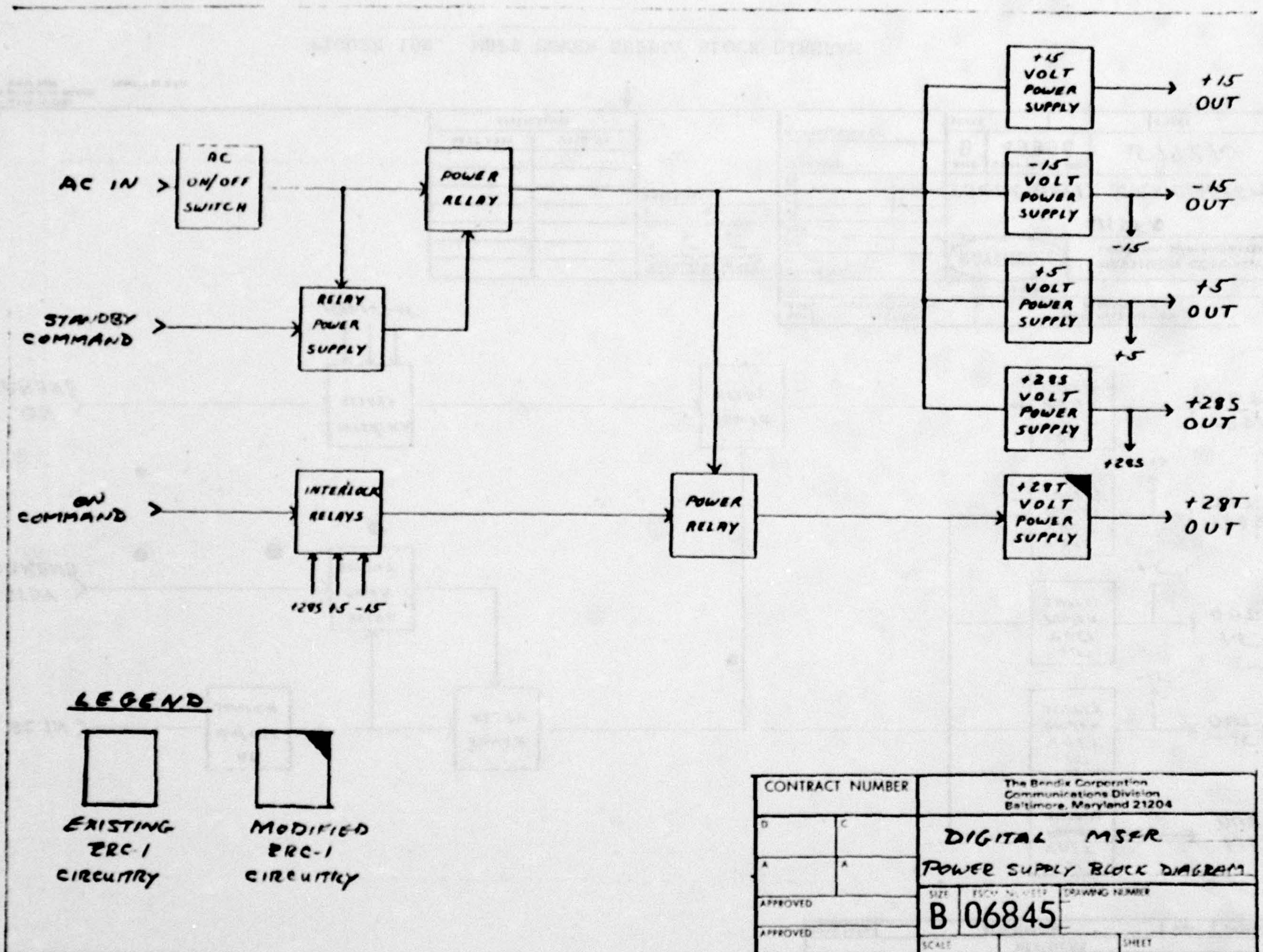


FIGURE 10A. DIGITAL MSFR POWER SUPPLY BLOCK DIAGRAM

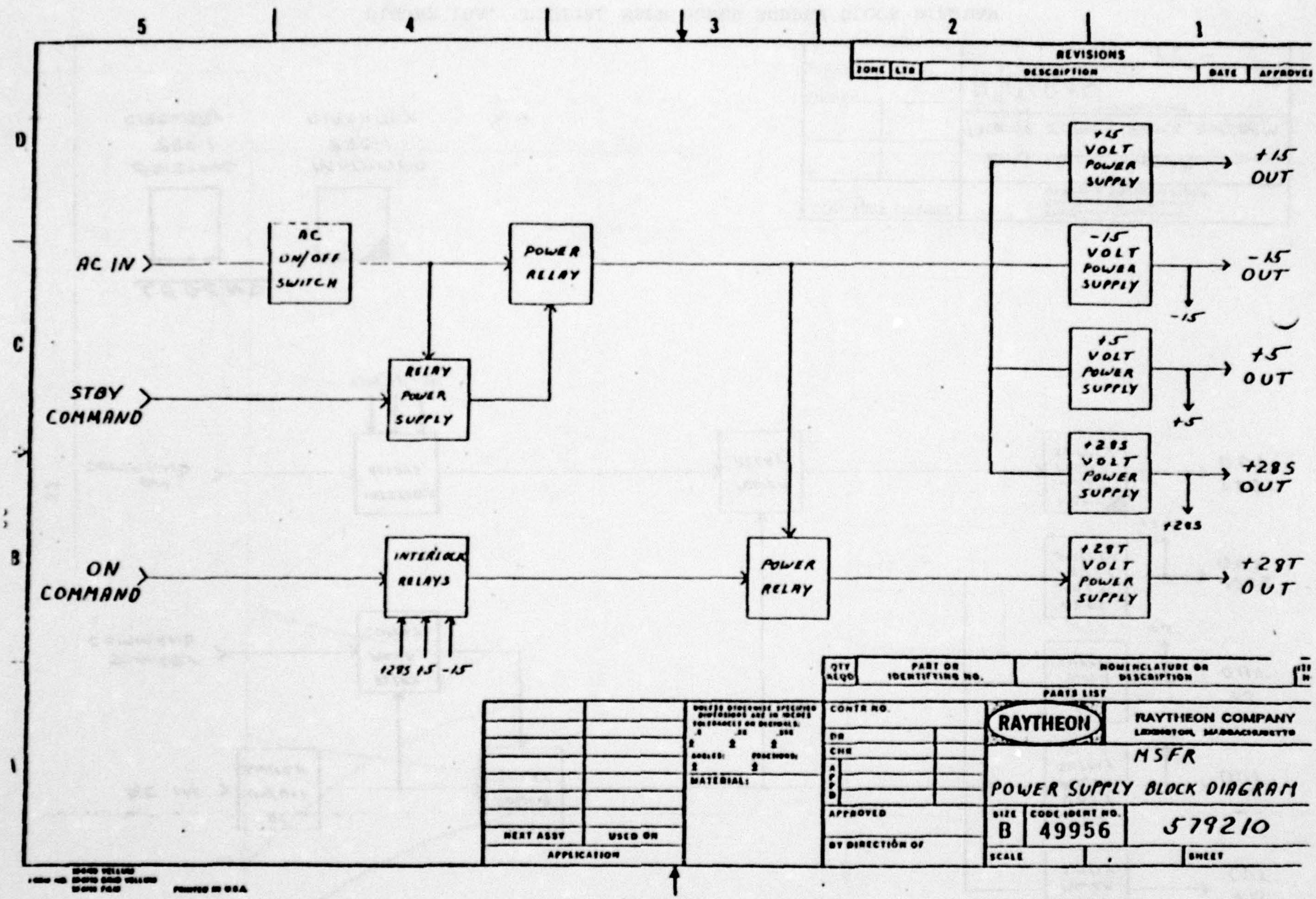


FIGURE 10B. MSFR POWER SUPPLY BLOCK DIAGRAM

APPENDIX A

BACKSCATTER ANALYSIS

The problem of backscatter due to the repeater's transmitted signal is considered. The analysis indicates that backscatter is a serious problem for the MSFR concept.

Consider a transmitted signal envelope as shown in Figure A1a. This would be the envelope of the signal transmitted by an MSFR with a 50% sampling duty cycle. Figure A1b shows how this waveform might look after being returned via terrain backscatter. This data was taken from simulation data found in a Raytheon report (AFAL-TR-69-327) and seems to correlate with empirical results found elsewhere in the literature[1,2]. Note that the return is spread out over several data bits. In fact, it has been found that the bulk of the returned energy for most terrains is somewhat non-coherent with respect to the transmitted signal. Note also that the returned signal is also relatively strong (for the case shown where h = 5000 ft).

---

1. Edison, A. R., et al, Radar Terrain Return Measure at Near-vertical Incidence, IRE Trans. on Antennas and Propagation, May, 1960
2. Ooms, J., Stargec, F. J., Corporate IR&D Study on Advanced SFR, Project Number 54413, 30 September 1966

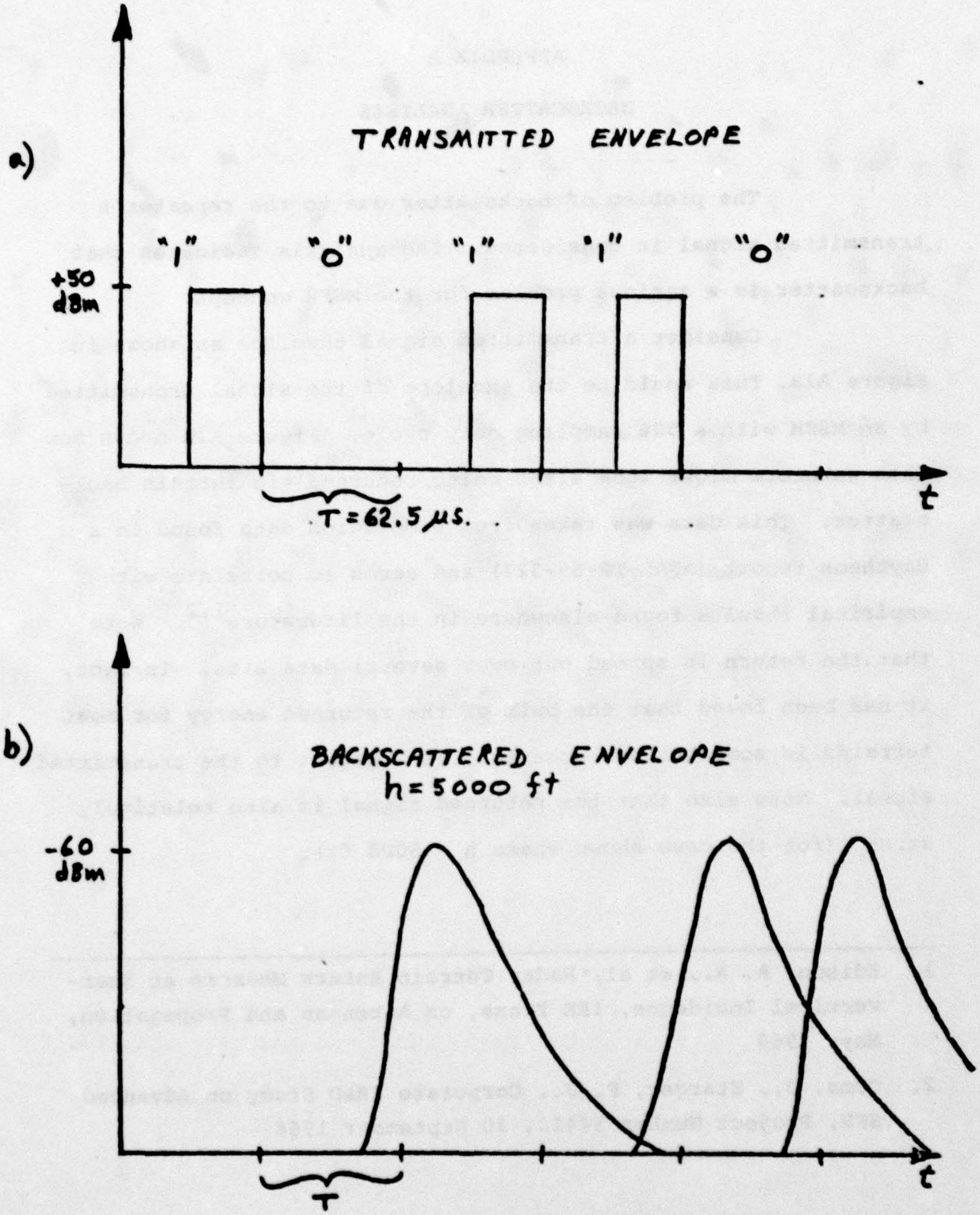

a)
TRANSMITTED ENVELOPE
"1" "0" "1" "1" "0"
+50 dBm
t
T=62.5 µs.
b)
BACKSCATTERED ENVELOPE
h=5000 ft
-60 dBm
t
T

FIGURE A1

A regressional analysis was performed on the backscatter data presented in the Raytheon report. The intent was to find a useful relation that could be used to estimate the path loss (= A_b) of the backscatter signal at the repeater's receiver for a given repeater height (h). Another relation was found to approximate the delay τ_b between the leading edge of the MSFR transmitted and the peak of the returned signal. It was found that

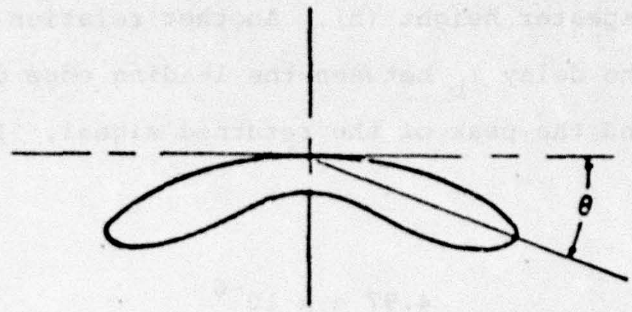
$$A_b = 110.7\, e^{4.97\, h \times 10^{-6}} \quad \text{dB} \qquad (1)$$

and

$$\tau_b = 35\, e^{4h \times 10^{-5}} \quad \mu\text{seconds} \qquad (2)$$

These equations represent average values one might expect. The data was averaged over many types of terrain and could vary as much as 10 or 15 dB (for A_b). In any event equations (1) and (2) provide a convenient basis to predict the orders of magnitude of A_b and τ_b. The antenna pattern used by Ratheon (Report AFAL-TR-69-327) to take the data from which equations (1) and (2) were derived is shown in Figure A2. Note that it provided some attenuation to backscattered signals with vertical incidence. Thus A_b would actually be slightly lower than predicted by equation (1).

If the MSFR is to be capable of receiving signals from users within some given radius, their received signal power at the repeater's receiver must be greater than the back-



θ	G
10°	-1.6 db
20°	0 db
40°	-1.3 db
60°	-6 db
80°	-7.3 db

FIGURE A2. PATTERN FOR BACKSCATTER PREDICTIONS scattered energy due to the repeater's transmitter. In fact, for an AM communication link such as that found with an ARC-164/MSFR link, the desired signal energy should be from 10 to 15 dB greater than the backscattered energy. Assume the line-of-sight path loss for the desired signal to be

$$L = 36.6 + 20 \log \frac{f}{MHz} + 20 \log \frac{d}{miles} \qquad (3)$$

If we let f = 400 MHz then

$$L = 88.64 + 20 \log d \qquad (4)$$

where d is the slant range between the transmitting ARC-164 and the MSFR. The ground range is given by d′ where

$$d' = d^2 + \frac{h^2}{(5280)^2} \qquad (5)$$

d = miles

h = feet

Assuming equal transmitted power and a 10 dB advantage is desired for the desired signal over the backscattered signal, equations (1), (4) and (5) combine to give

$$110.72\, e^{4.97\, h \times 10^{-6}} > 98.64 + 10 \log \; d^2 + \frac{h^2}{(5280)^2} \qquad (6)$$

Figure A3 is a plot of equation (6) in parametric form. Note that the left hand side of equation (6) represents the back-scatter path loss and the right hand side represents the desired signal's path loss. As long as this inequality is satisfied, the MSFR and the desired ARC-164 should be able to communicate.

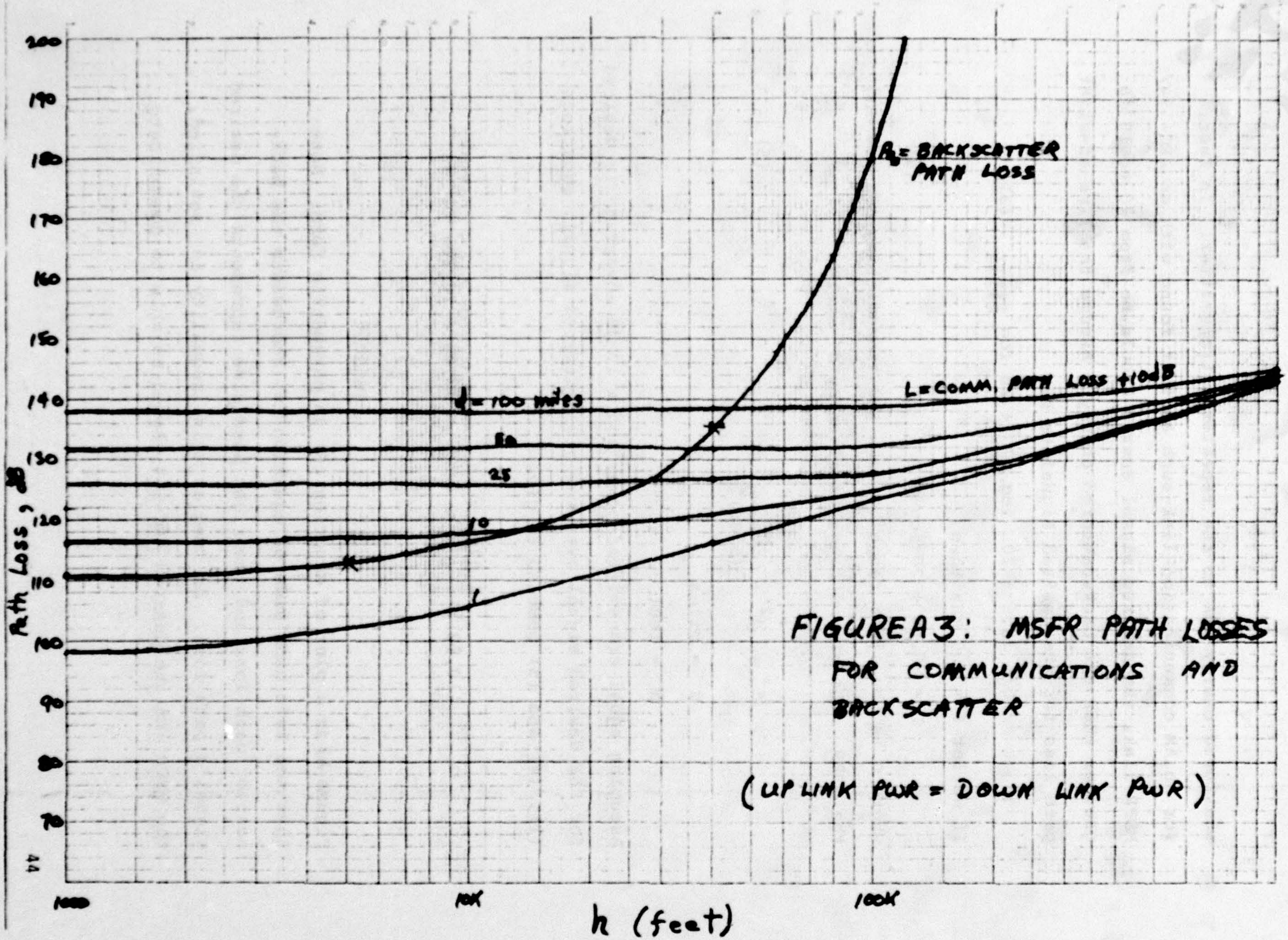


FIGURE A3: MSFR PATH LOSSES FOR COMMUNICATIONS AND BACKSCATTER

(UP LINK PWR = DOWN LINK PWR)

As an example, note that for communications path distances of 50 miles, the MSFR would have to be over 35,000 ft above the terrain (assuming an antenna pattern such as that shown in Figure A2).

In order to allow the repeater to be flown at lower altitudes, some form of backscatter cancellation scheme would be needed. The concept of adaptive cancellation has received a great deal of attention in recent years. Figure A4 shows how such a scheme might look if used to reduce the effects of backscatter in the MSFR. A pilot signal, which is orthogonal to the desired signal, is added to the desired signal. Thus any path taken by the desired transmitted signal back to the MSFR's receiver is also taken by the pilot. A portion of the transmitted signal and pilot is also passed through a complex weight and subtracted from the received signal. Only the pilot and the final residual received signal are compared to form the complex weights. This provides a means to cancel components of the received signal which contain the pilot. In theory, cancellation of the pilot should result in the cancellation of the transmitted signal at the receiver's input. In the case of the MSFR, though, it is felt that this scheme would not provide adequate cancellation of the backscattered waveform. The complex weight simply attenuates and delays a portion of the transmitted signal. Thus, at any given instant only one frequency can be fully cancelled. All other frequencies would get through to the receiver to some extent. Since backscatter delays can extend over several bits and appear as rather non-coherent energy it is felt that even several adaptive cancellers would not work well.

The design of an adequate pilot signal is also very difficult for the secure MSFR. The KY-58 is very sensitive to signals which look like 16 K baud digital data and is quick to reject any other signal structures. An adequate pilot signal must be orthogonal and thus look very different from the KY-58 signal. When this signal is added to the 58's waveform, it is likely that it would effect the operation of the KY-58 by sometimes causing the false dismissal of valid signals.

Rather than a single or even several complex weights in an adaptive cancellation scheme, it appears that a very complex system would be needed to adequately cancel the backscattered energy. The long delays (and wide range of delays) would indicate some type of long dispersive delay line with many programably weighted taps. It is still doubtful that even this would be a viable approach. The dynamic environment would mean that the delay line taps would have to be updated at an extremely fast rate and unless they are updated all at once, their effectiveness would always be at some suboptimum level.

It therefore looks as though the best way to avoid excessive backscatter with the secure MSFR would be to require that the repeater be flown at high altitudes. As long as the slant range is approximately equal to the ground range, equation (6) can be manipulated to give the minimum altitude

$$h \geq 2 \times 10^5 \ln(.891 + .18 \log d) \qquad (7)$$

where h is in feet and d is in miles. Although the same information is available from Figure A4, equation (7) is plotted for convenience in Figure A5.

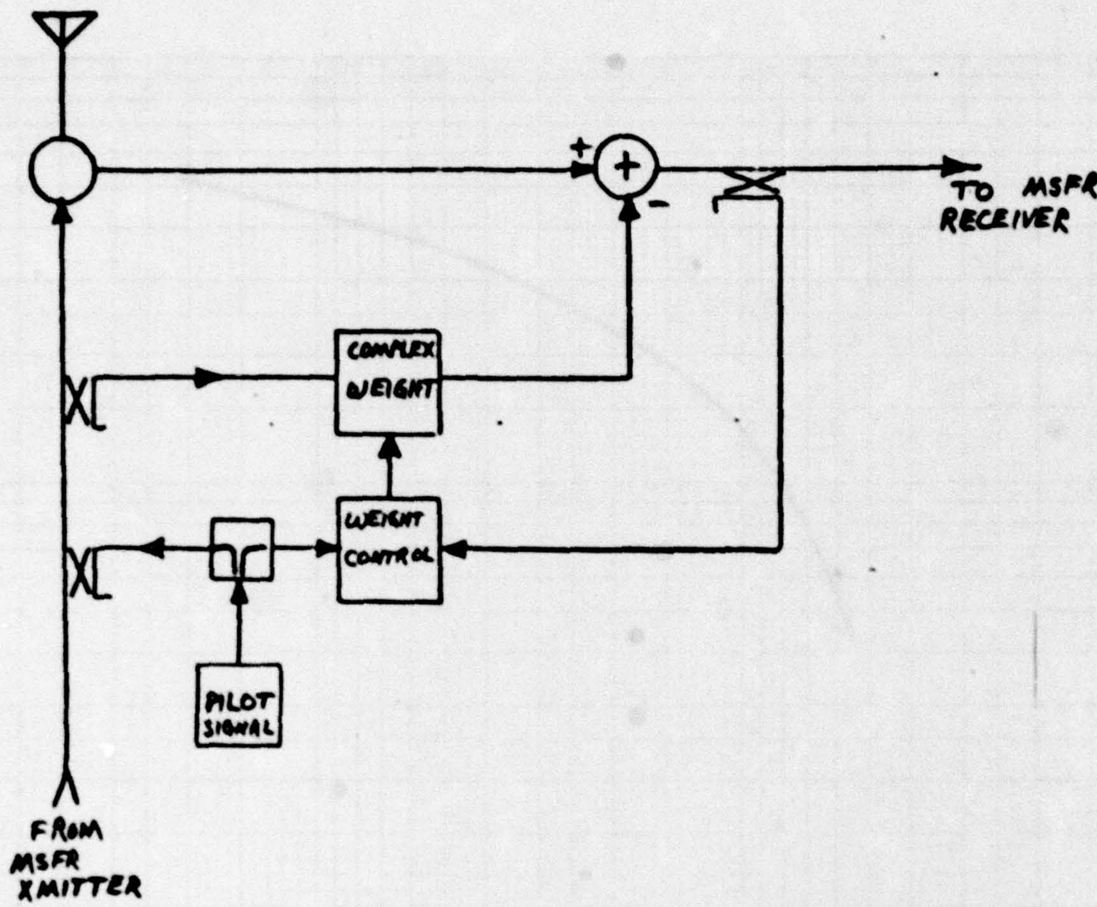
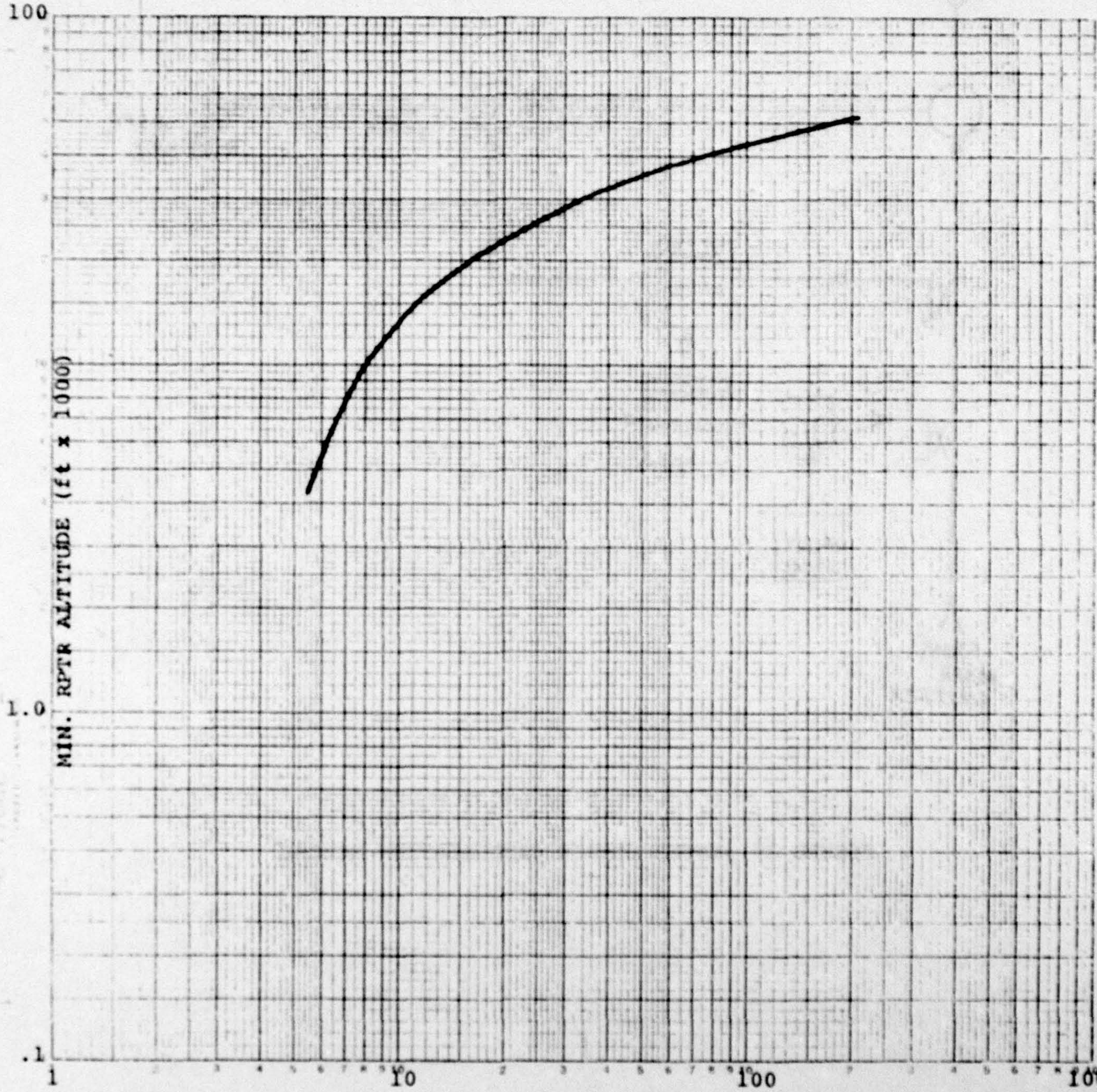


FIGURE A4. INTERFERENCE CANCELLATION SCHEME



COMM. RANGE (MILES)

FIGURE A5. MINIMUM REPEATER ALTITUDES

APPENDIX B

MSFR MULTICOUPLER

1.0 PURPOSE

The purpose of this Appendix will be to analyze the Digital MSFR's channel spacing and allocation requirements as a function of transmitter related broadband noise and intermodulation signal levels. Two technical analyses are performed; the first predicts the performance without transmitter or receive RF filtering while the second predicts the performance when a 4-channel RF multicoupler is used to attenuate off channel undesired signals.

2.0 INTRODUCTION

The proposed Digital MSFR operates from a single antenna as does its analog counterpart, the ZRC-1. However, operationally, the digital MSFR differs from the ZRC-1 in that one or more receivers and transmitters can operate simultaneously. Because of this characteristic and the inherent performance limitations of solid state transmitters, several undesirable performance characteristics, affecting channel spacing and assignment, need to be assessed. Two of these undesirable characteristics are broadband noise and Nth order intermodulation signal levels. The following section analyzes the expected performance of the digital MSFR relative to channel spacing and assignment with and without additional transmitter/receiver RF bandpass filtering.

3.0 TECHNICAL ANALYSIS

From ZRC-1 Specifications:

Transmitter

RF Output Power	+43 dBM
Noise Floor	-135 dBM
Third Order Intermodulation Conversion Coefficient	-25 dB

Receiver

Sensitivity*	-95 dBM = > 10 dB $\frac{S+N}{N}$

* -95 dBM is the typical sensitivity of the proposed Digital MSFR. It is calculated as follows.

ZRC-1 Sensitivity for 30% carrier modulation	-92 dBm
Correction for deletion of receiver input splitters	- 6 dB
Correction for 35 kHz receiver input splitters	+11 dB
Correction for multicoupler filter loss	+ 2 dB
Correction for 90% modulation of carrier	-10 dB
Net Digital MSFR Sensitivity	-95 dBm

3.1 DIGITAL MSFR CHANNEL SPACING AND ALLOCATION WITHOUT TRANSMITTER OUTPUT/RECEIVER INPUT FILTERING

Assume, for analysis purposes, that lossless hybrid signal splitters, having 20 dB output port isolation are used to combine the repeater receiver inputs and transmitter outputs. The broadband noise power generated by a single transmitter at the input to any receiver, from ZRC-1 specifications is -135 dBm/Hz. Considering that the Digital MSFR has an effective detector bandwidth of 35 kHz, the equivalent noise power at the input to the receiver would be -135 dBm/Hz + 10 log 35 kHz = -90 dBm. Therefore, to obtain a 0 dB S+N/N ratio, a -77 dBm 90% amplitude modulated carrier would be required. This corresponds to 18 dB of sensitivity degradation relative to a Digital MSFR receiver having a -91 dBm[1] (90% modulated) threshold sensitivity. In addition, if three transmitters are operating, almost 23 dB of sensitivity degradation would result. It is therefore apparent that transmitter output bandpass filtering is needed to restore the repeater's receiver sensitivity.

When 2 or more transmitters are operating on different frequencies, and combined using broadband combiners, Nth order spurious signals are generated. For the case of the Digital MSFR operating without transmitter output bandpass filtering, the frequency and level of 3rd order intermodulation products can be calculated as follows:

---

[1]Typical sensitivity of Digital MSFR configured with input hybrid splitters instead of multicoupler.

Assume for analysis purposes, a repeater is operational on the following frequencies 5 MHz apart.

Receiver, R1 operating on F1 of 295 MHz.

Transmitter T2 operating on F2 of 300 MHz.

Transmitter T3 operating on F3 of 305 MHz.

Receiver R4 operating on F4 of 310 MHz.

Transmitter output power of 43 dBM = P_o

Transmitter Combiner isolation 20 dB = I

Transmitter 3rd order intermodulation conversion coef. - 25 dB = C

3rd order inband intermodulation products are generated at 2F2-F3 and 2F3-F2; or 295 MHz and 310 MHz respectively. These products fall on F1 and F4, the receive channels. The signal level of the products $F1_{int}$ and $F4_{int}$ are calculated as follows:

$$F1_{int} = F4_{int} = Po + C + I \quad B =$$

$$= +43 \text{ dBM} -25 \text{ dB} -20 \text{ dB} -2 \text{ dBm},$$

where -25 dB is the transmitters 3rd order intermodulation conversion coefficient and -20 dB is the input port isolation of the power combiner. This extremely large level will require receiver frequency assignments many channels removed from the intermodulation frequencies. The actual spacing will depend upon receiver performance characteristics such as dynamic range, cross modulation, intermodulation, and input selectivity.

In addition, 5th and higher order products are generated on other frequencies although at somewhat reduced levels. Considering also that a third transmitter may be operating, additional products will be created further restricting channel assignment.

It is apparent that to avoid spectrum pollution and an exotic frequency management plan, transmitter RF output bandpass filtering is required.

3.2 DIGITAL MSFR CHANNEL SPACING AND ALLOCATION WITH A 4-CHANNEL MULTICOUPLER

Assume that each of the repeaters transmitter outputs and receiver inputs are filtered by an RF filter having the following attenuation characteristics.

3 dB bandwidth	>	2 MHz
60 dB bandwidth	$\leq$	10 MHz
80 dB bandwidth	$\leq$	20 MHz
insertion loss	$\leq$	2 dB

Transmitter boradband noise 5 MHz away from a receive channel is now attenuated 60 dB or to a level of -90 dBm -60 dB = -150 dBm, well below threshold sensitivity signal levels even with three transmitters operating. Channel spacings of 5 MHz would be conservative if only the effects of broadband noise had to be considered.

Nth order intermodulation product signal levels are dramatically reduced by transmitter filtering. Assume again the same repeater transmit and receive frequencies defined in section 3.1. The 3rd order intermodulation signals are calculated as follows:

$F1_{int} = F4_{int}$ = Po + C + 2 (Filter attn.@ 5 MHZ Δ)

+2 (Filter Insertion Loss)

= -43 dBm -25 dB -120 dB -4 dB

= 106 dBm

where

+43 dBM is the transmitter output power

-25 dB is the 3rd order intermodulation conversion coefficient

60 dB is the attenuation of the undesired signal at the desired frequency attributable to the desired signals filter. The 2x multiplier takes into account the fact that the desired channel filter has a symetrical attenuation vs frequency characteristic and the intermodulation product generated occurs at a frequency where it is attenuated by another 60 dB relative to the desired signal frequency.

The signal level is 11 dB below the -95 dBm threshold sensitivity of the Digital MSFR and should cause less than 1 dB of degradation. 5th and higher order products will be reduced even further due to typically lower transmitter intermodulation conversion coefficients and additional filter selectivity.

4.0 CONCLUSION

The Digital MSFR requires transmitter output filtering to achieve 5 MHz channel spacing and freedom from using frequency management. The proposed filter configuration would be a 4-channel multicoupler with an integral combiner. Each filter would be comprised of 4 poles having attenuation characteristics similar to those specified in section 3.2.

APPENDIX C

ANCILLARY PROCESSOR

1.0 INTRODUCTION

Oneof the Digital MSFR's operating characteristics is that each channels transmit/receive functions are time multiplexed during each VINSON Bit. This characteristic results in transmission of a unique digital signal to transceivers within the network. Although the transceivers demodulate the signal with little degradation, the signal provided to the COMSEC equipment for decryption is not useable without further processing.

This processing is accomplished by means of an ancillary processor that is relatively simple electrically and mechanically. It is incorporated in the signal path between the transceivers cipher text output and COMSEC equipments cipher text input.

2.0 ANCILLARY PROCESSOR TECHNICAL DESCRIPTION

Fig.C1 shows a circuit level functional block diagram for the ancillary processor. Operation is described below.

The Cipher Text (CT) signal from the transceiver is buffered and further filtered before being applied to squaring and integrate and dump circuitry. The squarer's function is to provide the phase detector with a signal having discrete 16 kHz components. The phase detector functions conventionally whereby the squarer's signal is compared to the 16 kHz reference signal. An error signal is generated that phase locks the VCXO frequency to the VINSON clock information. The phase locked VCXO signal is then used, after phase normalization, to properly time the

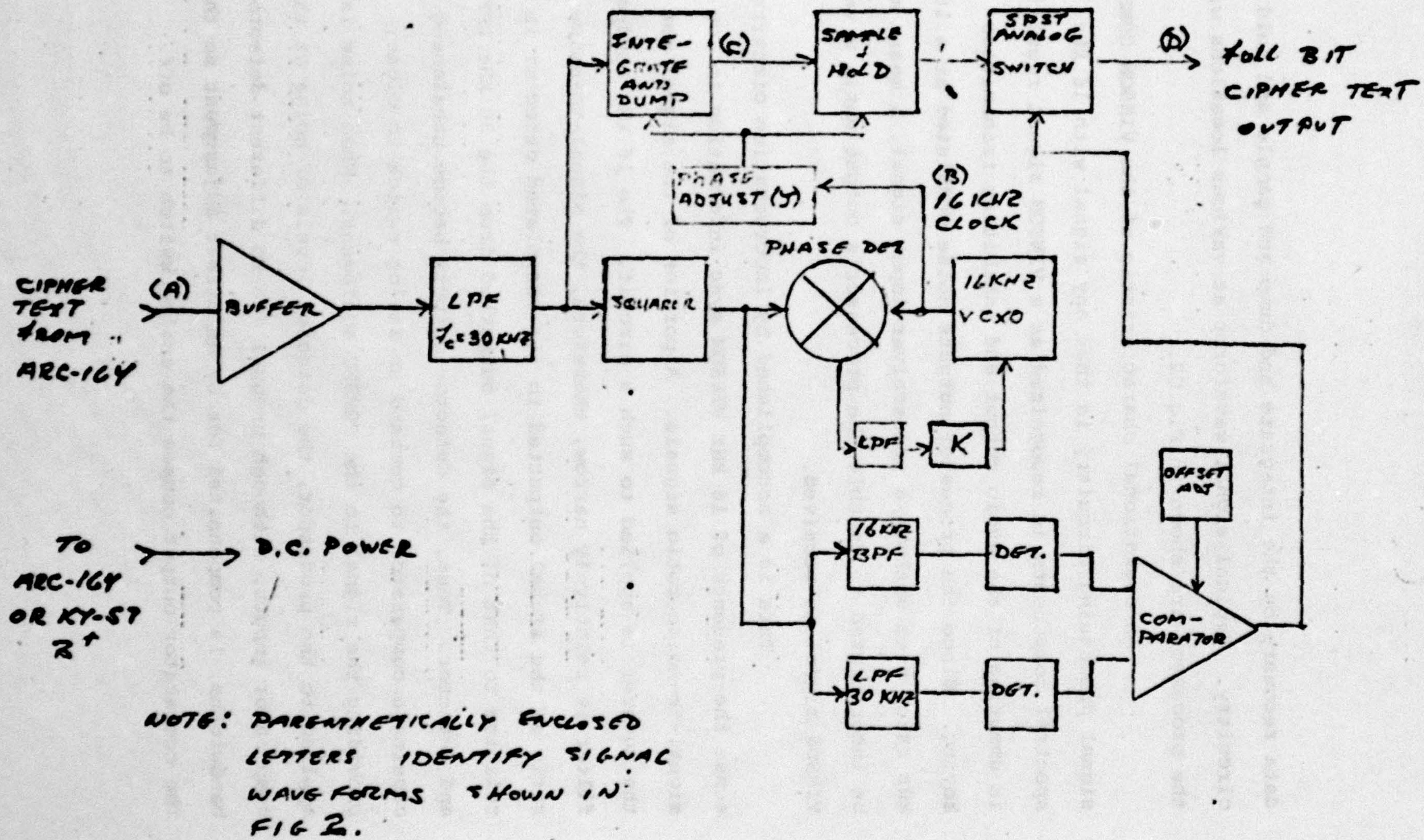


FIGURE C1. BLOCK DIAGRAM OF ANCILLARY PROCESSOR data recovery by the integrate and dump and sample and hold circuitry. Typical signal waveforms at various locations within the processor are shown in Fig.C2.

An operational characteristic of the VINSON COMSEC signal processing circuitry is that any signal with 16 kHz spectral components is recognized as a VINSON signal resulting in unmuting of the audio output and inhibiting transmitter keying. Since the processor outputs noise modulated at a 16 kHz rate, even without a transceiver input signal, a means must be incorporated to disable the processor's output until a valid VINSON signal is received.

This is a accomplished by incorporation circuitry to sense the presence of 16 kHz VINSON sync information in low signal-to-noise ratio signals. A portion of the signal out of the squarer is applied to such a circuit. The 16 kHz bandpass filter is relatively narrow, therefore, the signal-to-noise ratio of the signal outputted to its associated detector is high relative to that of the signal outputted from the 30 kHz LPF and detector. Thus, the detector outputs become unbalanced causing a comparator to command an analog switch to close, providing the signal to the COMSEC equipment. When noise is applied to the processor, the detected levels of noise at the comparator inputs, although unequal due to different detector bandwidths, is compensated for by an offset adjustment so that the comparator output causes the analog switch to be off.

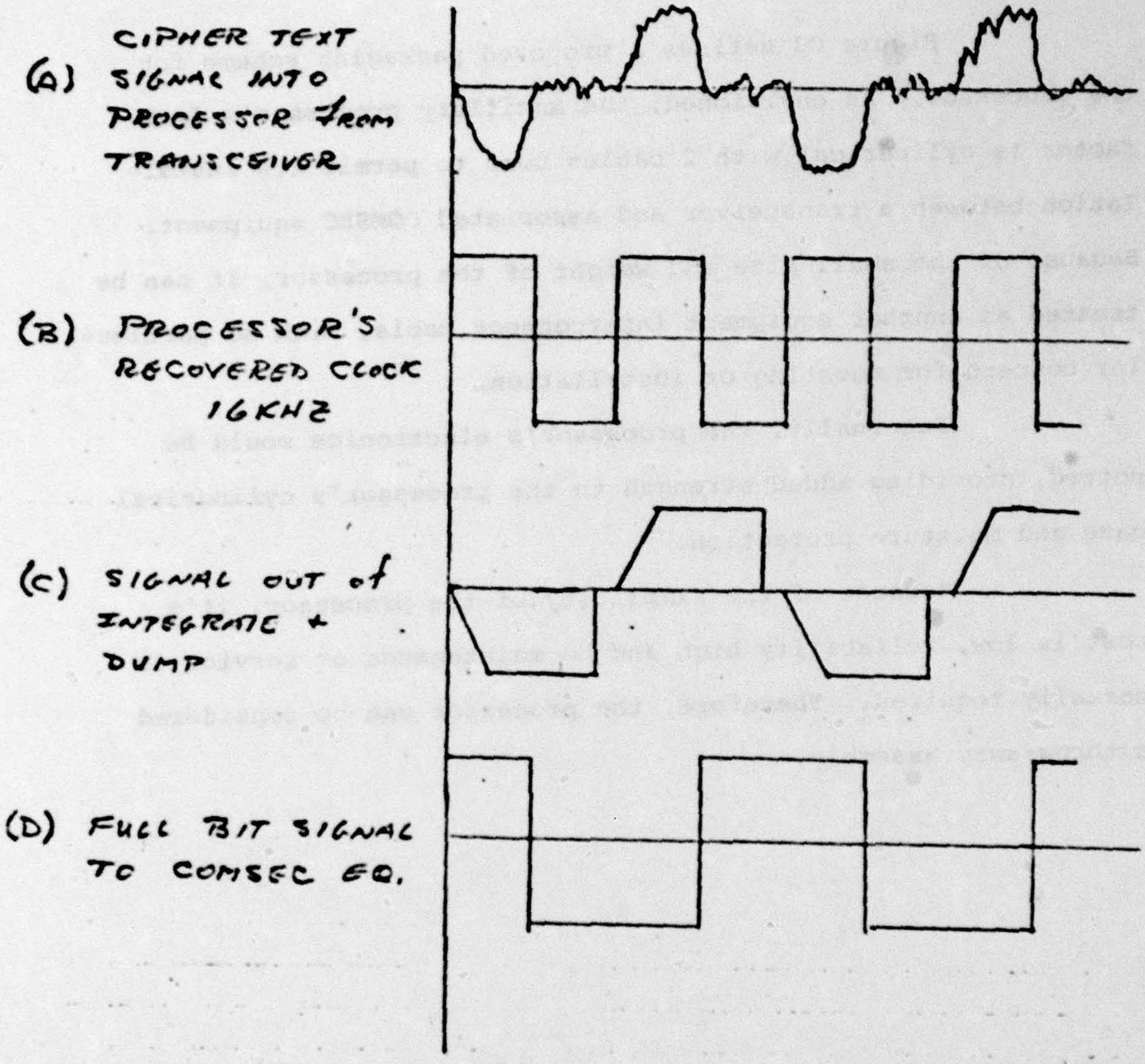


FIGURE C 2. TYPICAL ANCILLARY PROCESSOR WAVEFORMS

Figure C3 defines a proposed packaging scheme for the processor. As envisioned, the ancillary processor's form factor is cylindrical with 2 cables used to permit its installation between a transceiver and associated COMSEC equipment. Because of the small size and weight of the processor, it can be treated as another equipment interconnect cable, with no particular concern for mounting or installation.

Internally, the processor's electronics would be potted, providing added strength to the processor's cylindrical case and moisture protection.

Because of the simplicity of the processor, it's cost is low, reliability high and no maintenance or service is normally required. Therefore, the processor can be considered a throw-away assembly.

PROPOSED PACKAGE FOR ANCILLARY PROCESSOR

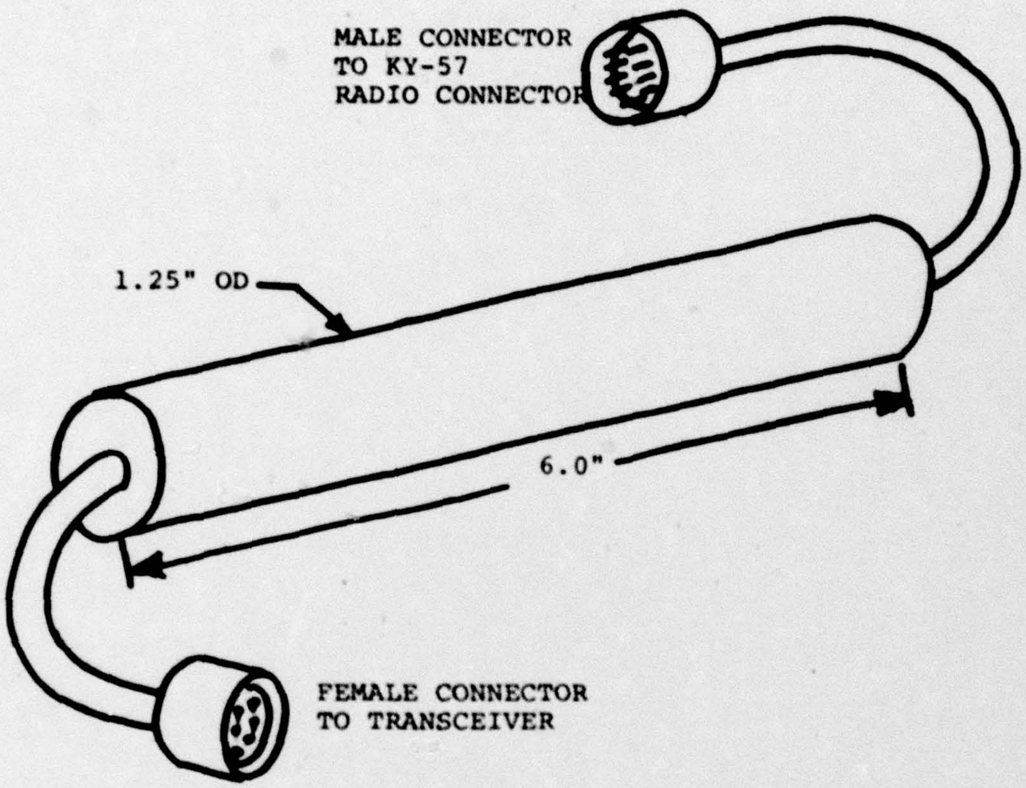


FIGURE B3